THE ELECTRONICS MAGAZINE WITH THE PRACTICAL APPROACH

UK £3·20 IR £4·70 (incl. VAT) July/August 1989

16 20 24 31 40 SUPPLEMENT
34 38 39 ONWARDS

Elektor Electronics

52 56 60 63

SUPPLEMENT OFFERING A VARIETY OF CONSTRUCTION PROJECTS

- Tracking tester
- Multi-layer PCBS
- Video compressor
- Floppy disk monitor
- Test pattern generator
- Digital model train (5)
- Practical filter design (7)
- RAM extension for BBC "B"

07
9 770268 451005

Elektor Electronics

July/August 1989
Volume 15
Number 169

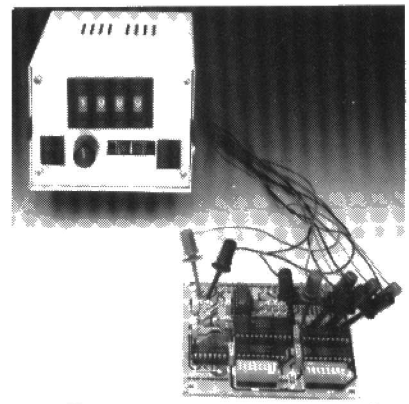
CONTENTS

Theme of the month in September will be Radio Techniques.

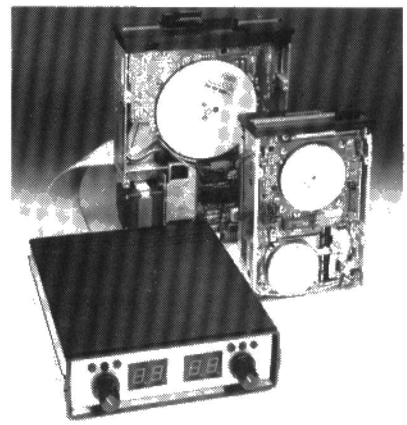
Also in the September issue:

- Centronics monitor
- ASIC microcontrollers from Intel
- Stereo viewer
- Communications receiver front-end filtering
- A high-grade power unit
- A new generation of analogue switches
- Simple transmission line experiments

Front cover
Our front cover this month illustrates the team work required in electronics design and development projects. The photograph was taken in our international design laboratory in the southern Netherlands.

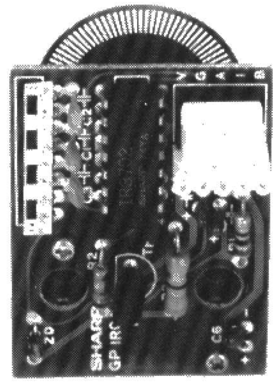


Test pattern generator – p. 16



Floppy disk monitor – p. 24

DC-AC power converter – p. 49



Optical shaft encoder – p. 60

ELEKTOR ELECTRONICS (Publishing)
Editor/publisher: Len Seymour
Technical Editor: J Buiting
Editorial offices:
Down House
Broomhill Road
LONDON SW18 4JQ
England
Telephone: 01-877 1688 *(National)*
or +44 1877 1688 *(International)*
Telex: 917003 (LPC G)
Fax: 01-874 9153 *(National)*
+44 1874 9153 *(International)*
Advertising: PRB Ltd
3 Wolseley Terrace
CHELTENHAM GL50 1TH
Telephone: (0242) 510760
Fax: (0242) 226626
European offices:
Postbus 75
6190 AB BEEK (L)
The Netherlands
Telephone: +31 4490 89444
Telex: 56617 (elekt nl)
Fax: +31 4490 70161
Director: M.M.J. Landman
Overseas editions:
Elektor sarl
Route Nationale; Le Seau; B.P. 53
59270 Bailleul — France
Editors: D R S Meyer;
G C P Raedersdorf
Elektor Verlag GmbH
Süsterfeld-Straße 25
5100 Aachen — West Germany
Editor: E J A Krempelsauer
Elektor EPE
Karaiskaki 14
16673 Voula — Athens — Greece
Editor: E Xanthoulis
Elektor Electronics PVT Ltd.
Chhotani Building
52 C, Proctor Road, Grant Road (E)
Bombay 400 007 — India
Editor: Surendra Iyer
Elektuur B.V.
Peter Treckpoelstraat 2-4
6191 VK Beek — the Netherlands
Editor: P E L Kersemakers
Electro-shop
35 Naseem Plaza
Lasbella Chawk
Karachi 5 — Pakistan
Manager: Zain Ahmed
Ferreira & Bento Lda.
R.D. Estefânia, 32-1°
1000 Lisboa — Portugal
Editor: Jeremias Sequeira
Ingelek S.A.
Plaza República Ecuador
2-28016 Madrid — Spain
Editor: A M Ferrer
Electronic Press AB
Box 5505
14105 Huddinge - Sweden
Editor: Bill Cedrum

Distribution:
SEYMOUR
1270 London Road
LONDON SW16 4DH.
Written and composed on Apple and IBM corporate publishing systems by Elektor Electronics.
Printed in the Netherlands by NDB, Zoeterwoude.

ABC
MEMBER OF THE AUDIT BUREAU OF CIRCULATIONS

HACKING: FUN OR EVIL?

There are indications that hacking, the nefarious accessing of computers, is becoming more widespread. There are many who in their naivity consider that this anti-social behaviour is nothing but a bit of fun whose perpetrators do a useful job in testing the security of computer systems. Nothing could be further from the truth.

A hacker is someone who, without permission, enters your home as it were and goes through your personal belongings. While doing so, he may, innocently or intentionally, take copies of your computer files. He may destroy or corrupt them. He may unwittingly plant an electronic time bomb that does not go off till much later.

There is no way of knowing exactly the harm and damage caused to private computer systems, but it is estimated, and in a number of countries a documented fact, that industrial and commercial losses world-wide amount to thousands of millions of pounds. More seriously, deaths have already occurred owing to the (hopefully inadvertent) changing of medical records by hackers infiltrating a French hospital computer system. And nobody knows how many cases of blackmail have been, or are being, conducted as the direct result of a hacker's activities.

It is, of course, true that many computer owners to a large extent have only themselves to blame. After all, you don't leave your car unlocked parked anywhere in a big (or even small) town nowadays. Nor do you leave the front door of your home open so that people can just walk in. That is, of course, no reason for anybody to take your car or enter your home, but it does hand them the opportunity of so doing on a plate. Considering the cost of a computer system, and the possible losses caused by hacking, the cost and trouble of making the system secure are relatively small.

Many may argue that the laws should be changed to deal with hackers, but that will not be the whole answer any more than it is in the case of other kinds of crime. Nevertheless, a change in the law is overdue, if only because it will at least reflect the attitude of the nation to hacking, an activity that is at best irresponsible, invariably harmful, and often destructive.

ELECTRONICS SCENE

Geographical Information System

A low-cost PC-based system, using videodisc technology, that opens up geographical information systems to a whole range of users and applications has been developed by Action Information.

'ActionPlan' is a hybrid system that plots video pictures and digital data together. The system is based on a 1000 km by 1000 km grid that allows simple, accurate and speedy placement of resources and rapid access to sites of interest.

At the heart of the system is the range of 'Maps in Action' videodiscs that provide background maps with up to 11 levels of magnification from a scale $1:10^7$ to very detailed maps with a scale of 1:1250.

UK's Most Powerful Chip

Plessey has produced the first micro-circuit to achieve the performance targets of a very high performance integrated circuit (VHPIC). The 1.4 µm double-layer metal CMOS chip contains 240,000 transistors, is about 1.1 cm², can be clocked at 30 MHz and performs 24-bit block floating point arithmetic.

Revolution in Telecoms

British Telecom is progressively implementing one of the world's most challenging computer projects to transform its service to customers. Known as Customer Service Systems (CCS), it is designed to draw together all the main elements of the customer services British Telecom provides. It has already been introduced in 13 of the 28 operating districts to cover more than eight million customers. By the end of this year, all 23 million customers will be able to experience its advantages.

CCS, the world's largest civil computing development, is part of British Telecom's £2,300 million-a-year investment programme to improve customer service and cater for growth. There is nothing comparable to it anywhere else in the world. Enquiries from Japan, the United States and several countries in western Europe indicate that it may well prove to be a significant foreign-currency earner.

Advanced Telecoms for Bangladesh Railways

Advanced telecommunications have been brought to Bangladesh Railways in Dhaka with the opening of the first phase of a £12 million 1700 kilometre optical fibre network.

The network, being installed by Telephone Cables Ltd (TCL), part of Britain's GEC Plessey Telecommunicasions Group (GPT), represents the first use of optical fibre technology in the country. It will be used for both telecommunications and train signalling, and also includes a microwave radio link across the 25 km wide River Jamuna in the north.

BS Approval for Siliconix JFETs

Siliconix has obtained BS CECC 50000 approval for two families of dual N-channel junction field-effect transistors (JFETs), the U430 and U431. These families consist of HF devices for use op to 100 MHz that have a low noise figure and guaranteed matching characteristics.

Transportable Radio Alarm Licensing Simplified

Operators of certain short-range transportable radio alarm systems will not need to be licensed individually under the Telecommunications Act since they will be covered by a class licence granted by the Department of Trade and Industry. Reference copies of this licence may be obtained for £1 from: The Library • Office of Telecommunications • Atlantic House • Holborn Viaduct • London EC1N 2HQ.

BT Takes Lead in Vocational Training

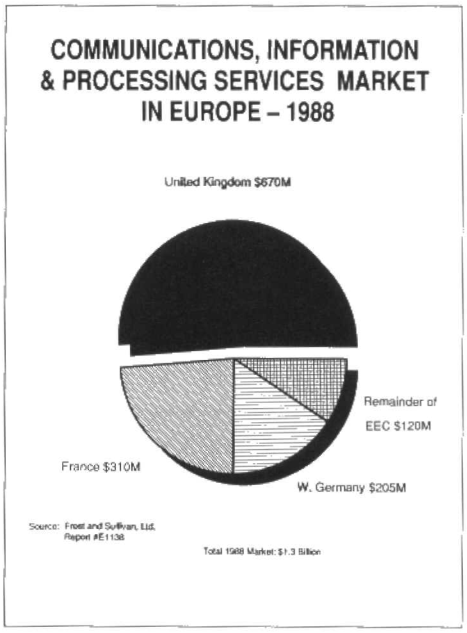
British Telecom has taken a lead in establishing vocational qualifications for the telecommunications industry. There are at present no industry-wide standards of competence.

British Telecom, together with STC Telecommunications, GEC Plessey Telecommunications, Cable & Wireless PLC, Mercury Communications Ltd and Telephone Rentals PLC, is setting up a body to act as a focus to determine standards of vocational competence throughout the industry.

Data Communications in Europe

According to a report* from Frost & Sullivan, the forecast for western Europe's data communications market looks very good.

Communications, information and processing services (CIPS) still constitute an infant market in Europe. In 1988, the market amounted to $1.3 billion, more than half of which in the UK. By 1993, Federal Germany will have gone ahead ($2.5 billion) with the UK second and France third ($2 billion).



* *A Strategic Report on European Data Communications* (#E1138)

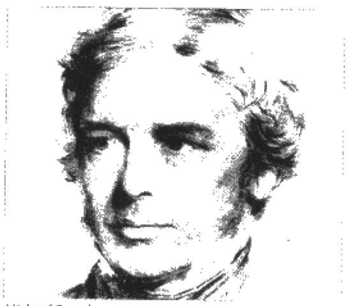
Bank of Scotland Presents the 1989/90 Faraday Lecture

Bank of Scotland has accepted the invitation from the Institution of Electrical Engineers to present the annual Faraday Lecture series. It is the first Scottish company – and the first financial institution – to undertake this exercise.

The lecture series, founded by the IEE in 1924, is used as a tribute to the memory of the scientist Michael Faraday, as a means of educating the general public of the significant strides being made in the area of Information Technology and to interest young people in career opportunities.

Recent presenters of the Faraday Lecture have been British Telecom, ICL, Ford Motor Company and the BBC. It is

in keeping with the Bank of Scotland's stance as a leader in the field of innovative banking that this opportunity has been accepted.



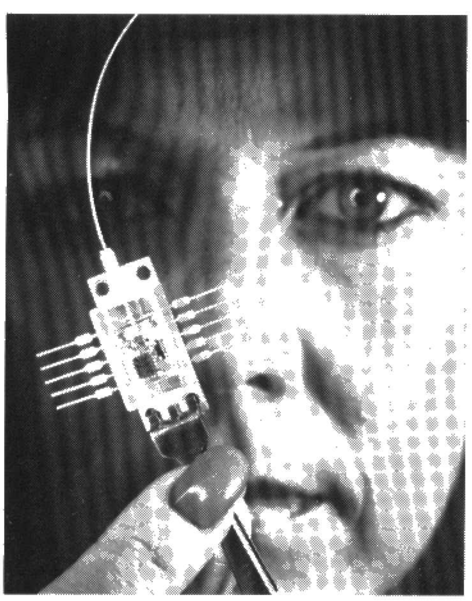
Michael Faraday: By courtesy of the Director of the Royal Institution.

Michael Faraday (1791–1867) pioneered the use of electricity and in 1831, by his discovery of electromagnetic induction, possibly his greatest achievement, he had invented the dynamo.

The lecture, entitled 'Electric Currency' and lasting for about an hour, will examine how electricity has reshaped the public's perception of money. It will be presented by a team of Bank staff at venues in 20 major cities throughout the UK, starting in Glasgow in September this year and finishing in Edinburgh in March 1990.

New Metal-coated Optical Fibre Allows Hermetically Sealed Connection

The Electro-optical and Data Systems Group of Hughes Aircraft Company has developed a metal-coated optical fibre that can be soldered to provide an hermetic seal resistant to a variety of harsh environments. The fibre, or pigtail, is used to connect an optical fibre cable to a package containing a laser or sensor and associated electronics as shown in the photograph below.



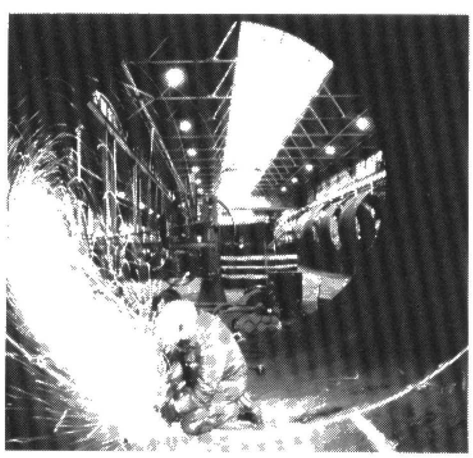
ELECTRONICS SCENE

Police Curb Noisy Bikes with New Device

The ear shattering noise from inadequately silenced motorcycles could soon be a thing of the past if a recent experiment continues to prove successful. Police in Derbyshire are using a hand-held sound level meter produced by Lucas CEL Instruments to advise bikers if their motorcycles exceed acceptable noise levels. During roadside checks, the meter is held half a metre from the rear of the exhaust at a 45° angle, and the engine run at half the total possible revolutions. If the exhaust noise exceeds 103 dB (which to most of us is still far too high), the rider is given a 'Vehicle Rectification Notice', which allows him/her seven days to effect repairs.

Educational Video on Noise in the Workplace

Draft regulations from the Health and Safety Commission on industrial noise, based on the EEC's 1986 directive, become law on 1 January 1990. By then, employers will need to know the facts about hearing damage, assess their own circumstances, and take the necessary action. Many employers who may consider themselves safe will find that lower action levels than recommended by the current code of practice will put them in a vulnerable position.



The importance and urgency of educating employers have prompted noise measurement specialist Bruel & Kjaer, Pitman Tutorial College, the BBC and the Institute of Sound and Vibration Research to co-operate in producing a video and accompanying text called 'Noise at Work – Action Now'

Major European Investment by US Combine

General Signal, a leading world-wide industrial manufacturer, is opening a new multi-million dollar centre for semiconductor manufactoring technology that aims to be the best equipped facility of its kind in Europe. The centre, which will incorporate General Signal's new European Headquarters for its Semiconductor Equipment Group, will be located in southern England.

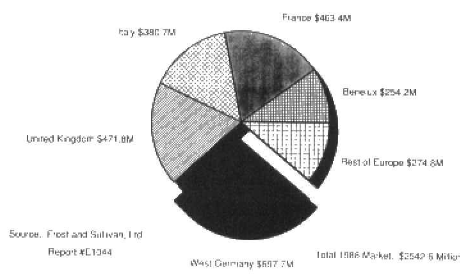
Sky Television Chooses GEC Mobile Radio Network

Sky Television has chosen the GEC National One communications network for the outside broadcast vehicles employed by camera and news crews.

Fibre Optic Telephone Lines Boost Power Supply Market

The change-over of telephone lines to fibre optics will help double the market for electronic power supplies in Europe say Frost & Sullivan in a recent report*.

SWITCHED MODE POWER SUPPLIES IN THE EUROPEAN MARKET - 1988



Although equipment from computers to TV receivers to X-ray machines needs a power supply, the report singles out telecommunications as the field of highest growth between now and 1993. It actually predicts that 18% annual increases will cause power supplies to eclipse computers and computer peripherals as the top demand area.

The United Kingdom will demonstrate a more rapid growth than will France or Germany. The latter is, at present, the largest market, however, with demand in 1988 reaching $698 million, that is, 27% of the total market. The UK market, at $472 million, represented 19% of the European total.

**Switched-mode Power Supplies in Europe* (#E1044).

BESSEL ARRAYS

Loudspeaker systems that use a relatively large number of identical drive units unfortunately provide a highly directional beam of sound. Depending on the alignment of the drive units, frequency 'holes' occur at a number of positions. Arranging the drive units in a Bessel array results in even sound distribution. It should be noted that for the commercial use of the systems described in this article the permission of the patentee, Philips Export BV, Eindhoven, the Netherlands, is required

The requirement for frequency-independent even distribution of sound in large rooms can be met only approximately. Radiation patterns at different frequencies but equal magnitude are shown in Fig. 1. From these diagrams, it is clear that the pattern approaches the ideal at low frequencies and that the higher the frequency the more chaotic the distribution. Assuming a maximum sound pressure of 85 dBm, the sound pressure in the 'eyes' of the 1600 Hz diagram is some 40 dB down, but only 8 dB at 100 Hz.

This problem is tackled by arranging a number of drive units in an array. If high powers are required, further difficulties are encountered, because the individual drive units then become relatively more expensive and less suited for the present purpose, since the large cones produce more distortion. The use of a number of smaller drive units connected in parallel or series also has disadvantages. One of these is that the more drive units are used and the higher the frequency is, the smaller the angle of radiation will become.

The distortion of the radiation pattern at higher frequencies is very troublesome and, for example, does not permit stereo radiation.

The Bessel array developed in the Philips research laboratories makes it possible to configure a number of identical drive units, without additional components, in a manner where they virtually behave as one entity with a higher power rating.

Bessel functions

The researches at Philips have shown that a spatially interconnected number of small, identical drive units has a near-spherical radiation pattern if the total power applied to them is divided over the individual units in accordance with a Bessel function. Each drive unit, m, is then powered according to the function:

$$J_m(p) = \sum_{k=0}^{k=\infty} \frac{(-1)^k \, (p/2)^{m+2k}}{k!(m+k)!}$$

For the first drive unit (m = 1), the value of the function is normalized at 1. With a suitable value for p, the functions $J_{m+1}(p)$, $J_{m+2}(p)$... become whole multiples of $J_m(p)$. This mathematical trick obviates fractions for the valency of the drive units.

The sum of all the functions is 1 and that is the reason that a number of small units in a Bessel array have the same radiation pattern as one large one. Strictly speaking, this is true only for an infinite array, but good results are achieved already when m = 5.

Simple Bessel series

Fig. 2 shows a series layout of five drive units separated by a distance d and two possible circuits that satisfy the function ratio of

A:B:C:D:E: = 1:2:2:–2:1.

The normalized power applied to units A and E is exactly half supplied to B and C. Unit D, which needs a power of –2, has its terminals reversed. The two circuits are only different in their resulting impedance. Assuming that all the drive units are 8 Ω, the impedance of the series circuit is 28 Ω and that of the parallel circuit is 2.3 Ω.

A similar set-up, but for seven drive units, is shown in Fig. 3. Here, the weighting factors are:

A:B:C:D:E:F:G = 1:2:2:0:–2:2:–1.

The factor 0 merely means that unit D is omitted: in other words, the distance between units C and E is $2d$.

When nine drive units are considered, some will have the factor 0. Their ratio is:

A:B:C:D:E:F:G:H:I = 1:2:2:0:–2:0:2:–2:1.

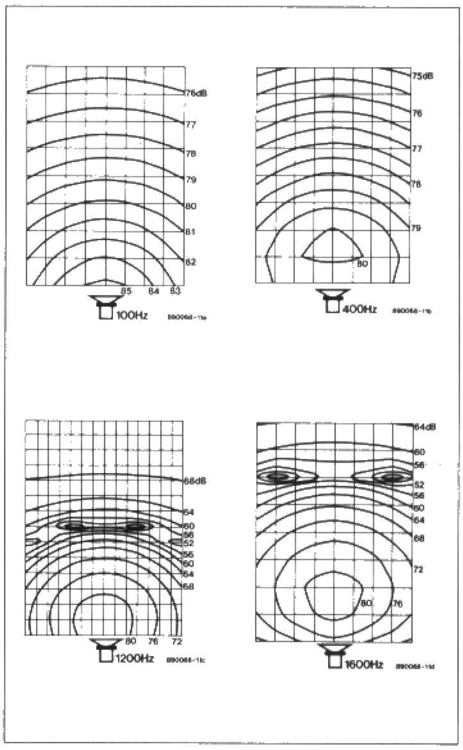


Fig. 1. Radiation pattern of a drive unit at different frequencies but at constant sound pressure.

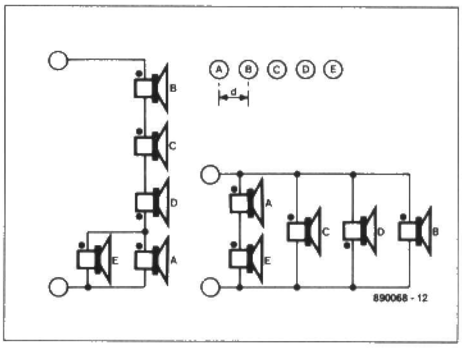


Fig. 2. Simple Bessel series with five identical drive units. Note the polarity of the units as indicated by the bullets. Distance *d* is only slightly larger than the loudspeaker diameter.

If your maths is up to it, you can compute higher orders. The procedure is to first normalize $J_1(p)$, that is, make it 1 and give a value to p.

It is, of course, not necessary to com-

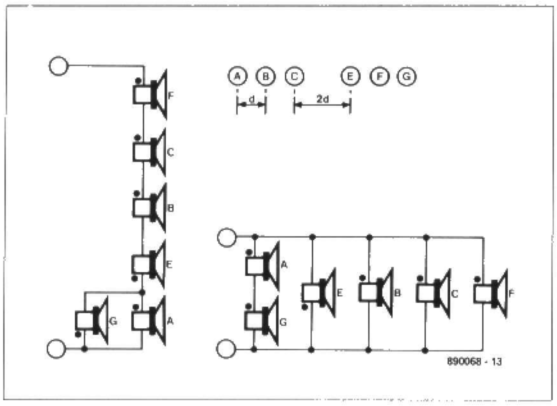


Fig. 3. A Bessel series with seven weighting factors needs only six drive units.

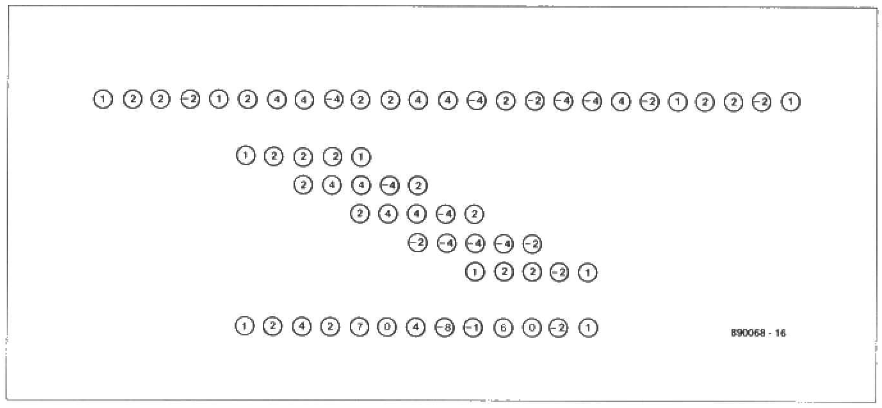


Fig. 6. Twenty-five drive units can be combined in a fairly simple manner (a). It is possible by rearranging the weightings (b) to reduce the number of drive units to 11.

pute the sum to infinity: the terms quickly become smaller and, depending on the value of p, may be neglected for $k > 6$.

Array combinations

Simple Bessel series may be combined. If a radial pattern in only one direction is required, a number of identical arrays may be combined as shown in Fig. 4. This configuration is suitable when the listener(s) is always in a horizontal plane perpendicular to the array surface as, for exam-ple, in a theatre or cinema.

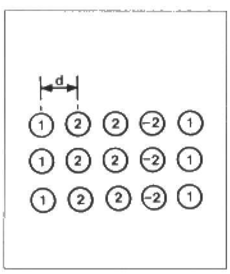


Fig. 4. Identical arrays placed side by side give a particularly horizontal radiation pattern.

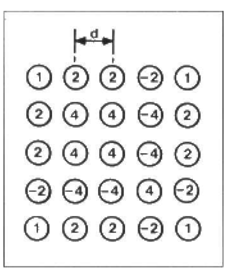


Fig. 5. A spherical radiation pattern requires identical rows and columns.

A different approach is necessary when a semi-spherical pattern is required, for instance, when the array is suspended from the ceiling, that is, parallel with the floor. It is then necessary to combine identical arrays both in the rows and the columns as shown in Fig. 5. The same effect is obtained with the configuration of Fig. 6a, comprising five groups of five drive units. The valency of each group is the same as that of the simple array in Fig. 2. Within each group, the valency is divided in the same way and by the same method.

It is, of course, not an easy matter to configure 25 drive units in a row. It is, however, amazingly simple to reduce the number, because it is possible to rearrange the weighting factors of Fig. 6a as shown in Fig. 6b and this results in only 13 factors and 11 drive units. How far the rows are displaced with respect to one another depends on the number of wanted drive units, the maximum power at the highest weighting, the maximum weighting ratio and some other factors. The result is, however, always the same.

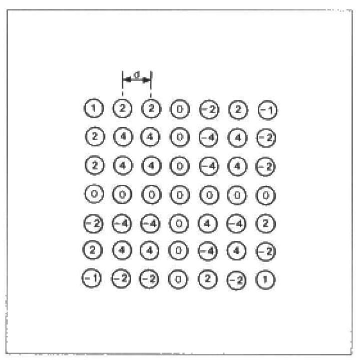


Fig. 7. A seven by seven Bessel array.

The configuration of 49 drive units in Fig. 7 is intended to satisfy the aesthetic requirements of interior decorators.

Stereo arrays

In stereo applications, it is not so that two Bessel arrays are required. If from the two channel outputs a sum signal (L+R) and a difference signal (L–R) are produced, these signals are simply connected to the five-factor drive unit configuration of Fig. 8. The weighting factor ratio is:

$$A:B:C:D:E = k(L+R):2(1-k)(L-R):$$
$$:2k(L+R):-2(1-k)(L-R):k(L+R).$$

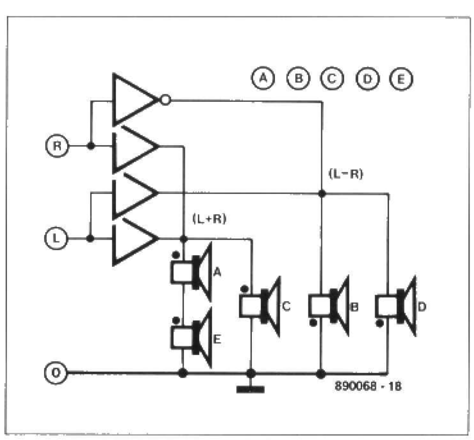


Fig. 8. Layout of the stereo version.

This ratio coincides with that of the mono version (test this by omitting the right-hand channel). The simple factor k offers many possibilities. For example, if the amplifier input is preceded by the circuit of Fig. 9, a continuously variable base width setting may be realized with the following values for k.

If $k = 0$, super stereo results; if k lies between 0 and 0.5, expanded stereo is obtained; $k = 0.5$ gives normal stereo as in Fig.8; and $k = 1$ results in mono operation.

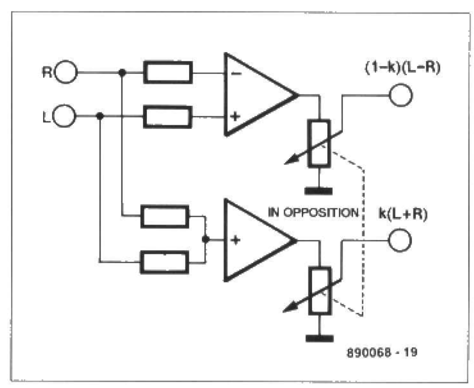


Fig. 9. Circuit to obtain a continuously variable base width setting.

It will, no doubt, be of interest to many readers that in the above the drive units may be replaced by microphones to yield a supersensitive microphone with a spherical radiation pattern.

Literature: *Bessel panels – high-power speaker systems with radiual sound distribution*. Philips Technical Publication 091.

TEST PATTERN GENERATOR

A. Rigby

A programmable digital pattern generator for all of you who do not have access to specialized equipment for faultfinding in digital circuits.

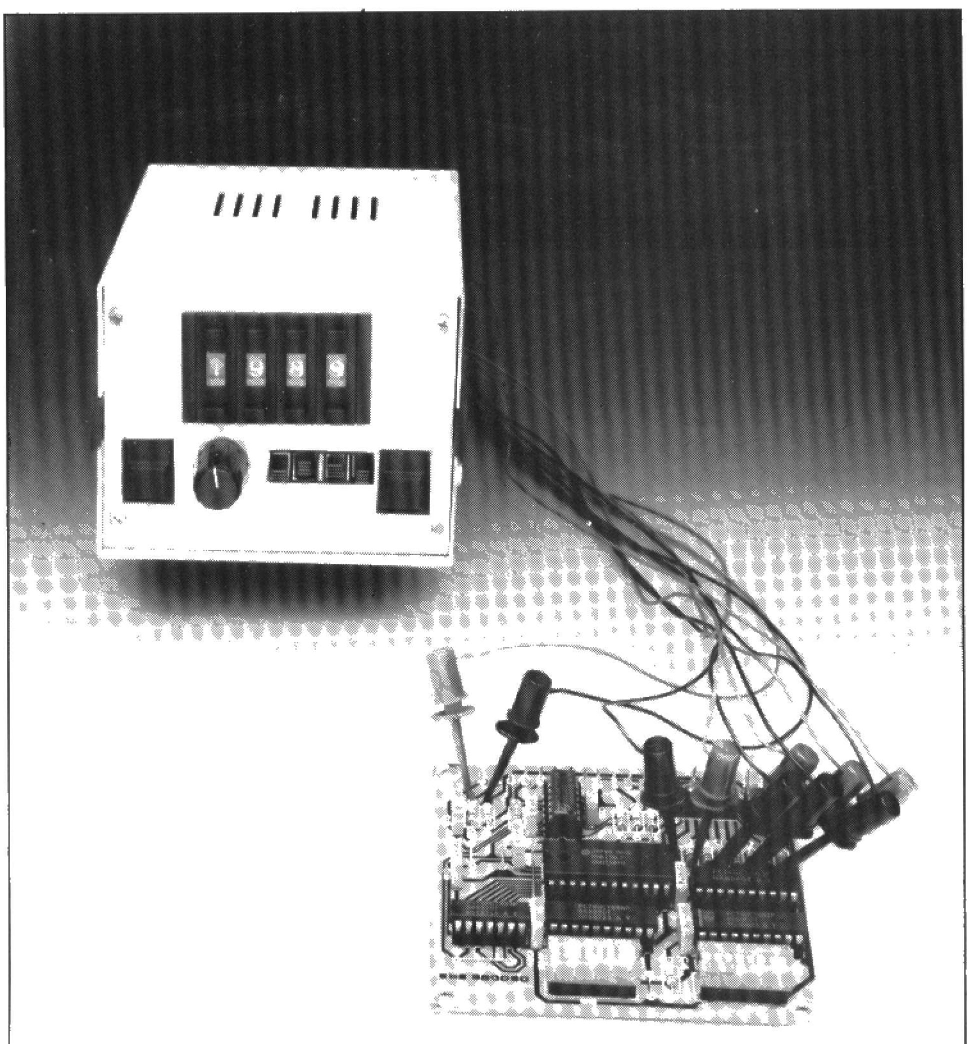
The operation of many digital circuits is, in principle, not simpler or more complex than that of analogue circuits, but the dependency of certain signals on others gives digital faultfinding a labyrinth effect. Furthermore, it is often impossible to isolate a particular section of the circuit for a stand-alone test. The omission of a single control signal, whose function may not be known at all initially, may cause the whole circuit to 'stall', making it impossible to track the cause of the malfunction.

In synchronously operating digital circuits, state changes take place only as a result of clock signal transitions. This makes the operation of this type of circuit relatively simple to follow, especially if the clock signal or signals can be generated by external means. By contrast, a level change at a particular point in an asynchronous digital circuit is taken over by the rest of the circuit. This means that a spurious pulse at any point in the circuit can easily disrupt the normal operation of the entire digital equipment.

A test pattern generator as described here enables programmed data to appear in a predefined order at the input of the circuit under test. An oscilloscope is used to check whether the circuit gives the correct response to the applied test patterns.

Patterns for testing

The generator is capable of supplying up to 255 8-bit test words, or a sequence thereof. The number of test words can be set by the user. A short sequence therefore takes hardly time to program, and need only be programmed once. A sequence of, say, five test words simply corresponds to five bytes loaded into the memory of the test generator. The remaining 250 memory positions are not used and need not be loaded.



The block diagram in Fig. 1 shows that the heart of the circuit is formed by a RAM memory. Two 8-way DIL switch blocks serve to program the test words and the length of the test word sequence. The memory locations are addressed by a counter, which counts from 0 to 255. The generated address (A) is compared to the preset length of the sequence (B). When A and B are equal, output A=B of the word comparator is actuated. Depending on the position of the mode selection switch, the test pattern is either stopped or repeated. When the switch is set to CONTINUOUS, the output signal of the comparator causes the address counter to be reset and to start counting from 0 again. When the mode switch is set to SINGLE, the oscillator is inhibited, so that the last counter state is 'frozen'.

There are three more switches in the circuit. One of these serves to select either the 'single step' or the 'run' mode. In single-step mode, the clock pulse for the circuit is generated by a push-button, while in run mode it is generated by means of a clock generator. Switch OC enables the outputs to be switched to high-impedance (three-state). The last switch, marked RUN/PROG causes the $\overline{WE}$ (*write enable*) input of the memory to be connected to the clock signal (program), or to the positive supply voltage (run). In the run mode, the memory is made 'read-only'.

Circuit description

The circuit diagram of Fig. 2 shows how the previously discussed functions are given their practical form. The test generator is composed of only six integrated circuits.

Since only 256 of the 2,048 memory locations in RAM IC3 are used, address lines A8, A9 and A10 are connected to ground. The Type 6116 2K×8 CMOS static RAM is used here because it is currently

cheaper and more readily available than any 256-byte type.

The counter that drives the eight remaining address lines of the RAM is formed by IC_2, a Type 74HCT393. This 8-bit bistable counts continuously from 0 to the value set with switch block S_8. IC_4 compares the current counter state at inputs Q_n with the preset word at inputs P_n. When the words are equal, output P=Q goes low. Depending on the position of S_4, either the outputs of N_3 and N_4 are blocked, or the counter is reset. After being reset, the counter starts at state 0 again.

The memory databus is connected to an input buffer with three-state outputs, IC_1, and an output buffer, IC_5. When the memory is read out, RAM input $\overline{WE}$ is logic high, and IC_1 is switched to high-impedance mode. The output buffer is transparent, and latches the data from the databus on the negative edge of the clock signal. The buffer passes all data applied to its D inputs as long as input C is logic high. When C is made logic low, the negative clock pulse transition causes the data at the D inputs to be latched in the buffer's internal registers. The data remains there and on the outputs, until input C is made high again. The $\overline{OC}$ input allows the output drivers in the buffer to be switched to high-impedance, so that the test generator is effectively disconnected from the circuit under test without the need of leads to be unplugged or wires to be removed.

Manual or automatically?

As already discussed, the circuit can be controlled either by a clock generator or with the aid of the SINGLE STEP switch, S_1. R-C network R_1-C_1 suppresses bounce pulses generated when the SINGLE STEP button is actuated. Schmitt-trigger gate N_3 gives the switch signal a digital shape and level.

Gate N_4, capacitor C_2 and resistors P_1 and R_3 form an adjustable clock generator. One input of N_3 and N_4 is connected to switch S_4, which enables the gates to be blocked. When S_4 is set to CONTINUOUS, one input of both gates is taken high via R_2, so that the clock oscillator is enabled. The span of potentiometer P_1 is fairly large at 1:100, allowing the user to set the optimum test frequency for many types of digital circuit.

External push-button S_6, with associated debouncing network R_9-C_3, makes it possible to reset the counter. The enable inputs of IC_1 and the $\overline{WE}$ input of IC_3 are logic high as long as S_3 is open. The input buffer is in high-impedance mode, and IC_3 behaves as a read-only memory. The circuit is thus set to the 'run' mode.

When S_3 is closed, input $\overline{WE}$ is pulled low during the active part of the clock pulse, and the input buffer is enabled. The circuit is in 'program' mode because the dataword set on S_7 is stored into the memory during the rising edge of the clock pulse. The clock signal at the pole of S_2 is applied to the clock input of bistable FF_1

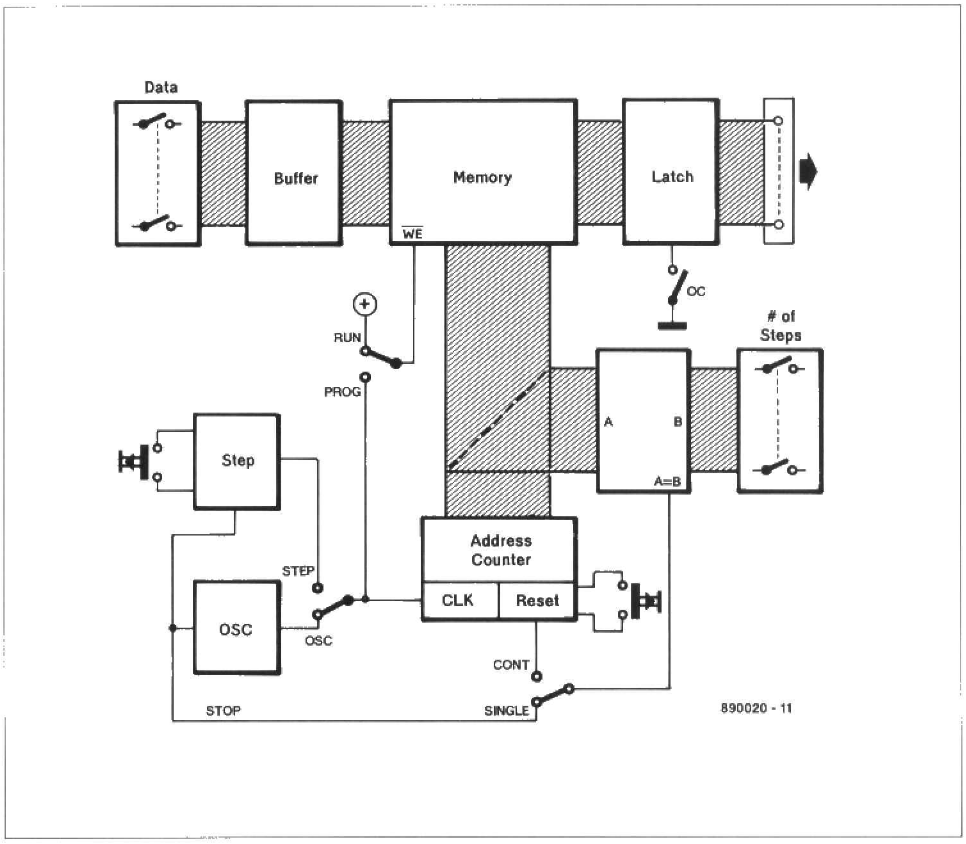


Fig. 1. Block diagram of the test pattern generator for digital circuits.

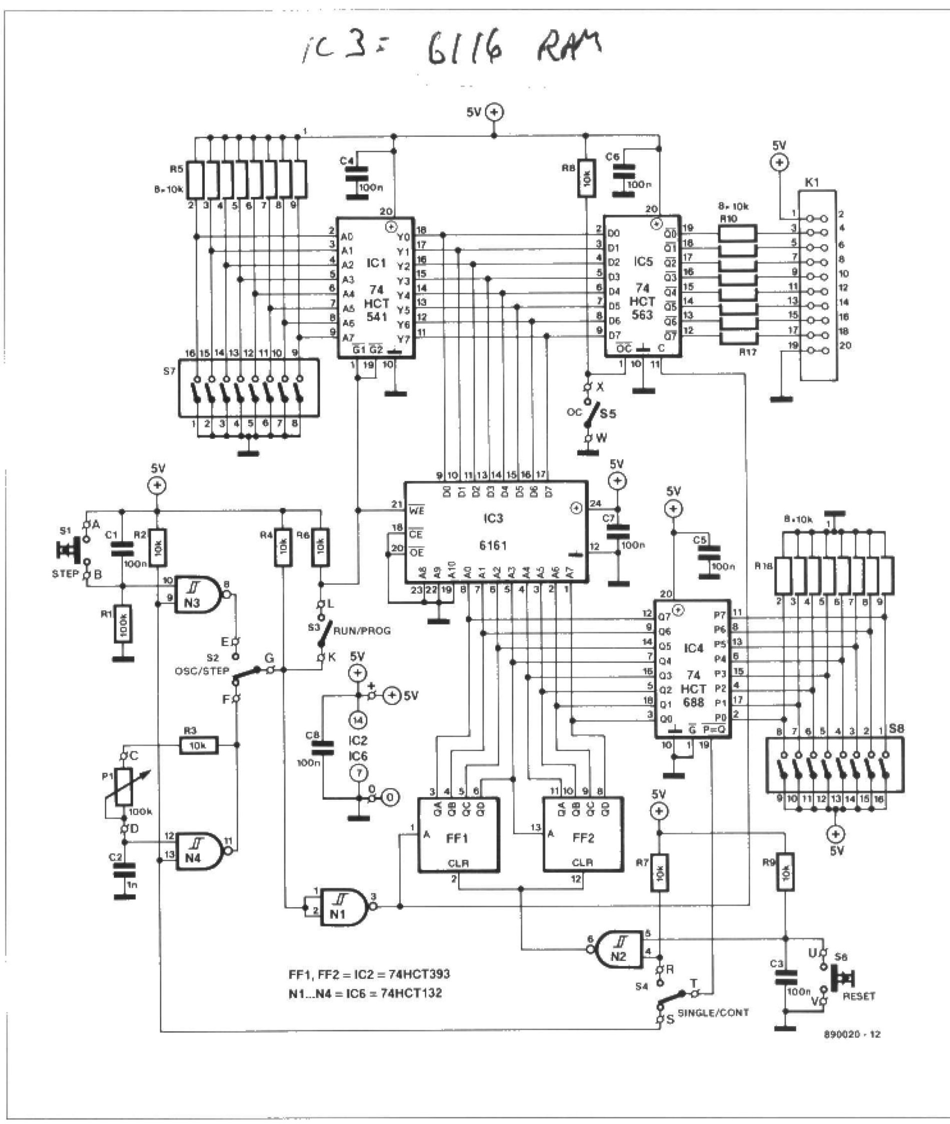
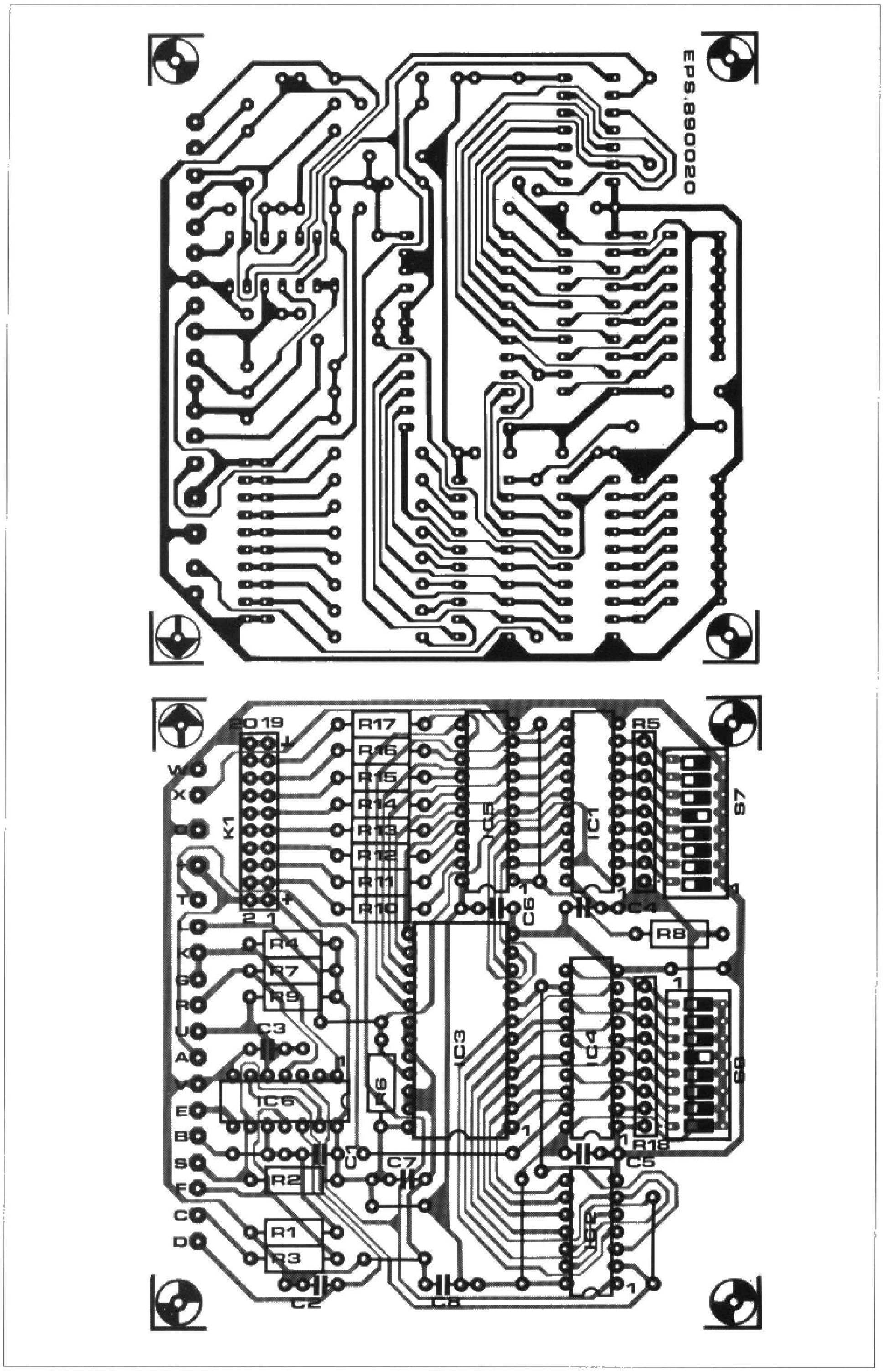


Fig. 2. Circuit diagram of the tester.



Parts list

Resistors (±5%):
R_1 = 100k
R_2;R_3;R_4;R_6–R_{17} = 10k
R_5;R_{18} = SIL resistor array 8×10k
P_1 = potentiometer 100k linear

Capacitors:
C_1;C_3–C_8 = 100n
C_2 = 1n0

Semiconductors:
IC_1 = 74HCT541
IC_2 = 74HCT393
IC_3 = 6116 or 8416
IC_4 = 74HCT688
IC_5 = 74HCT563
IC_6 = 74HCT132

Miscellaneous:
S_1;S_6 = SPST push-to-make button.
S_2;S_3;S_4 = miniature SPDT switch.
S_5 = miniature SPST switch.
S_7;S_8 = 8-way DIP switch or hex thumb-wheel-switch.
K_1 = 20-way pin header.
PCB Type 890020 (not available ready-made through the Readers Services).

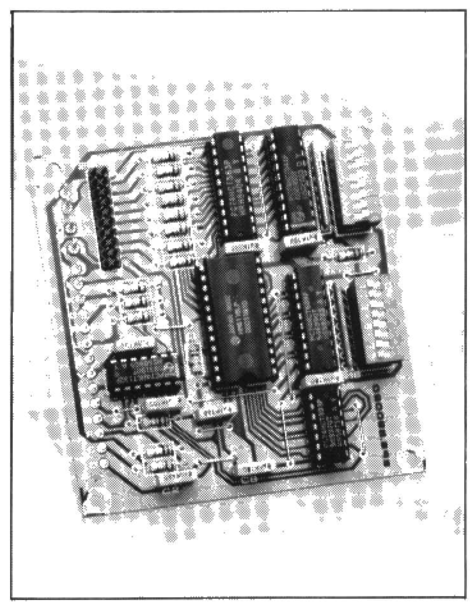


Fig. 3. Track layout and component mounting plan.

via inverter N_1. Output Q4 of FF_1 in turn clocks a second bistable, FF_2. The bistables (there are actually four in each device) are thus cascaded to form an 8-bit counter.

Suggestions

The clock frequency has been chosen rather arbitrarily but will be suitable for most applications. If required, the value of C_2 should be adapted to give a different frequency. A larger value of C_2 results in a lower clock frequency, and a smaller one in a higher clock frequency. It is also possible to create a larger frequency range by selecting different capacitors by means of a rotary switch.

Toggle switch S_2 may be replaced by a three-position type. In that case, the third position is used for selecting an external clock source, e.g., one available in the circuit under test.

Construction

The compact printed-circuit board designed for the tester makes construction a matter of routine. The track layout and component overlay are given in Fig. 3.

Start the construction with the fitting of the ten wire links on the board. Next, fit the 18 solder terminals and connector K_2. The HCT ICs are all low-cost types, so that sockets are not strictly required. Although the board allows the fitting of two 8-way DIL switch blocks, it is better, in many cases, to use switches that can be mounted on to the enclosure. Hexadecimal thumbwheel switches are convenient in the practical use of the test generator and are, therefore, suggested as an more ergonomical and simple-to-connect alternative to DIP switch blocks.

The power supply is purposely not accommodated on the board because a regulated 5 V source will nearly always be available as part of the circuit under test. The digital test generator draws about 30 mA.

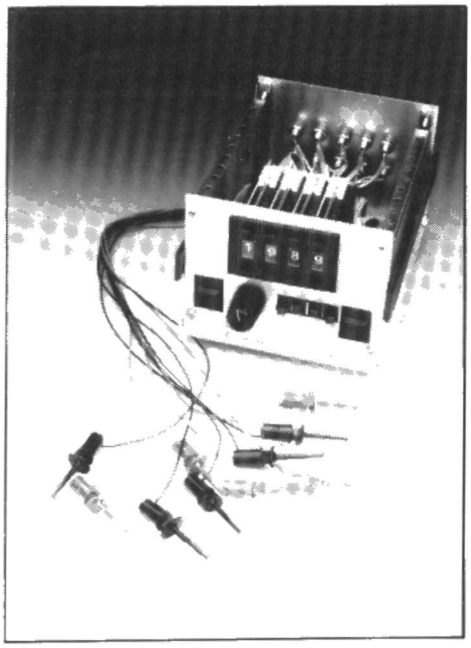


Fig. 4. Prototype of the test generator housed in a compact enclosure with the word configuration and word number switches mounted on to the front panel.

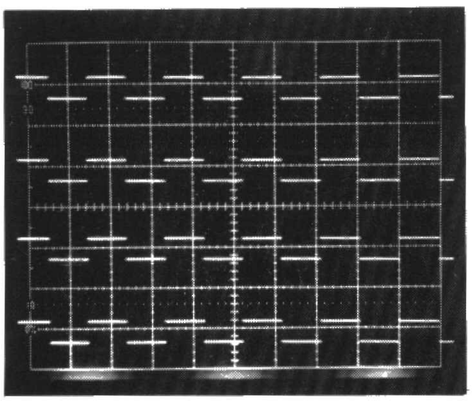


Fig. 5. Bit pattern displayed on a logic analyser.

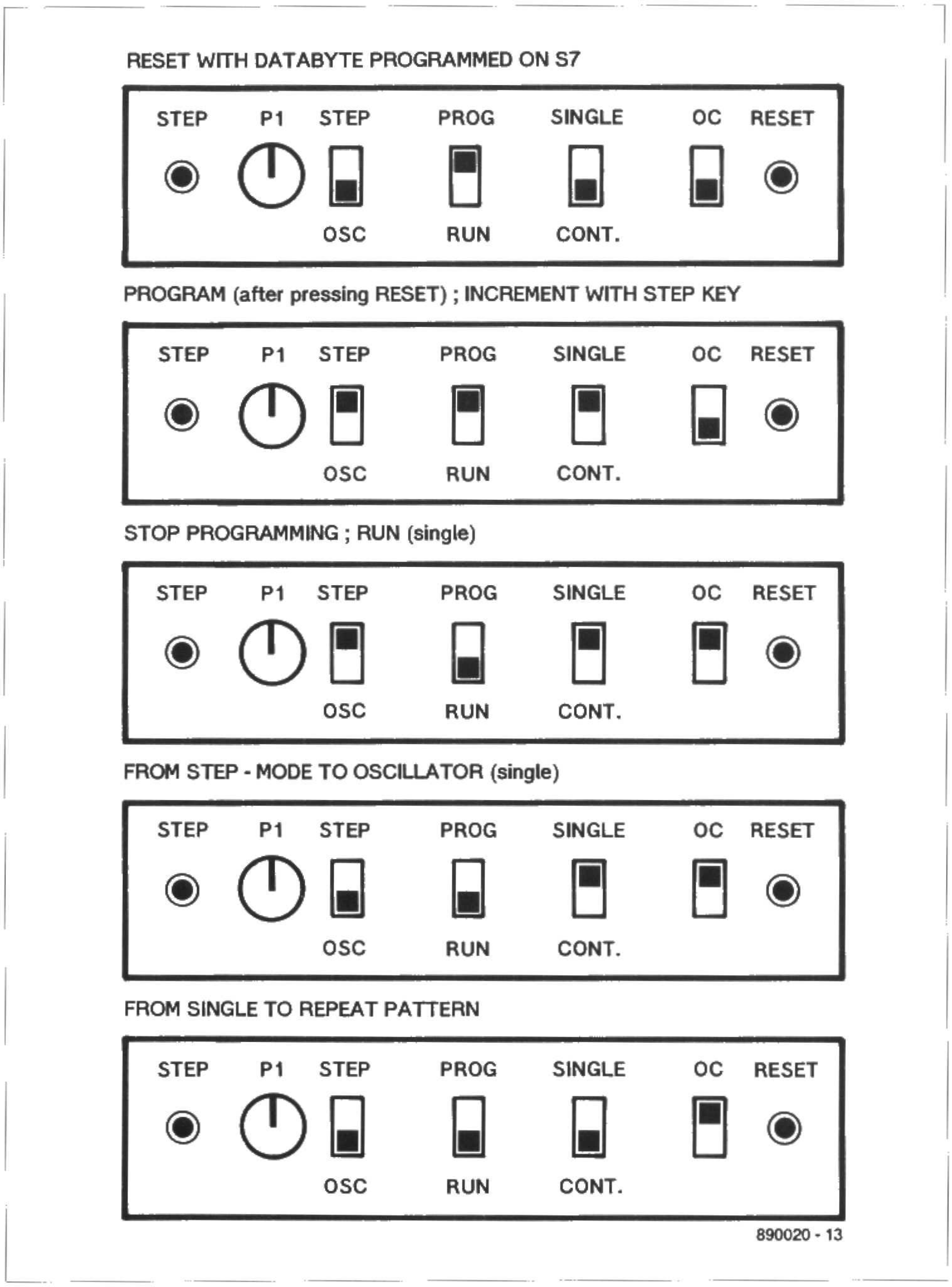


Fig. 6. Overview of the programming and loading operations using the front panel controls.

Programming

The memory has to be loaded with the desired bit-patterns (test words) before the circuit can be used to test a digital system. Fortunately, programming is straightforward:

- Set the number of test words as a hexadecimal value on S_8 (00_H–FF_H).
- Set S_3 to PROGRAM, S_2 to STEP and S_4 to SINGLE. Set the desired bit pattern on S_7.
- Press STEP to store the bit pattern in memory.
- Set the next bit pattern.

When S_4 is set to SINGLE, the circuit will not accept further data when the number set with S_8 is reached. After loading all bit patterns, S_3 is set to RUN to sequentially feed the data to the circuit under test. Depending on the position of S_2, this feeding out takes place automatically or manually.

Timing and level measurements may start after the digital outputs of the tester have been connected to relevant points in the circuit under test. The tester does not provide a strobe pulse, but this is fairly simple to implement by programming, say, bit 8 accordingly. In that case, the test word is 7-bits wide, and the sequential data stream has a maximum length of 128 samples because the strobe bit must toggle in between samples.

Finally, one interesting application of the test generator should not be left unmentioned here. It is fairly simple to use the instrument for driving a D-A (*digital-to-analogue*) converter. This combination creates a simple programmable waveform generator.

LINCMOS CIRCUITS

LinCMOS™ is a process that gives to linear devices a superior performance over metal-gate CMOS by the use of polysilicon gates and an optimized 'N well' structure. Equivalents of many popular operational amplifiers, comparators and timers have already been available for some time. The major benefits of these devices are lower power consumption, faster switching and the ability to operate from very low supply voltages.

While giving good ± supply rail performance, with a total voltage not exceeding 16 V, the input and output are optimized for single supply operation. This is achieved with an input common mode range that included GND ($-V_{DD}$ with ± supplies) and an output range that pulls down to within a few millivolts of GND (with a load connected to GND). The TLC27x range are specified to work with supply voltage down to 3 V and will thus operate with the supplies that are commony available for TTL and HCMOS. For maximum dynamic range, single rail operation with 16 V supplies should be used. For low power and battery operation, the TLC25x range are specified to operate with 1 V total supply voltage.

High bias mode gives a wider bandwidth (2.3 MHz) and faster slew rate (4.5 V/µs) than standard bipolar opamps (especially single-supply devices) for the same order of supply current. The enhanced bandwidth gives an increase in the open-loop to closed-loop gain ratio at a particular frequency improving accuracy of, for example, filter circuits, or allowing higher frequency operation. Slew rate enhancement gives a wider large-signal bandwidth and generally allows the implementation of faster circuits.

Medium bias mode gives standard bipolar opamp performance at roughly a tenth of the supply current.

The main advantage of the low bias mode is the low power consumption with sufficient bandwidth and slew rate for basic sensor interface and audio applications.

Low bias and offset currents allow circuit simplification through the elimination of bias current balancing resistors, higher impedance circuits for greater accuracy (for instance, smaller, higher tolerance capacitors) and circuit current defined only by feedback components. Another advantage is insignificant noise due to bias current (shot noise): noise is dominated by noise voltage and resistor noise—see Fig 3.

LinCMOS is a trade mark of Texas Instruments.

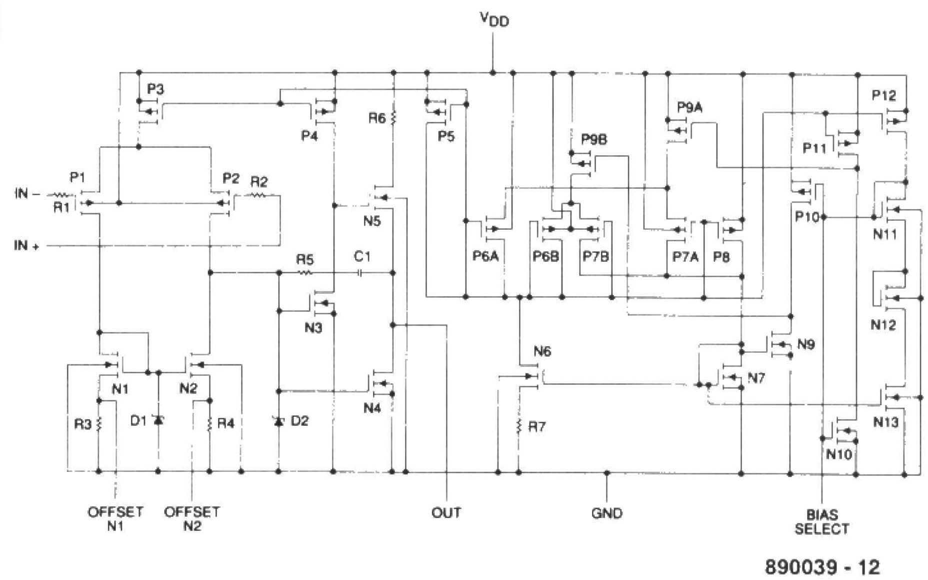


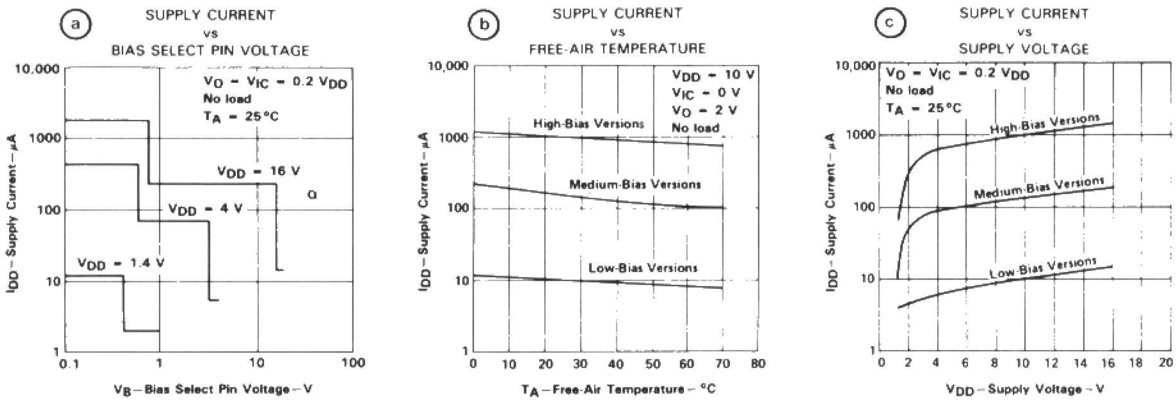
Fig. 1. Circuit diagram of a typical LinCMOS operational amplifier.

TECHNOLOGY	BIPOLAR[1]	BIFET[2]	LinCMOS (bias levels) LOW	MEDIUM	HIGH
V_{CC} MAX REC	±15	±15	16	16	16
V_{CC} MIN REC	±5	±5	1	1	1
V_{CC} SPECFD	±15	±15	10**	10**	10**
V_{IO} mV	1–10	3–20	2–10	2–10	2–10
αV_{IO} uV/degC	5–20*	10	0.7	2	5
I_{IO}	2–750nA	5p–2nA	1–300pA	1–300pA	1–300pA
I_{IB}	20–800nA	30p–10nA	1–600pA	1–600pA	1–600pA
V_{ICR} V	±13 or Vcc−1.5	±12	−0.2 to 9	−0.2 to 9	−0.2 to 9
V_{OM} IN 10kohms	24–26 or Vcc−1.5	24–77	0–7.8 RL=1M	0–7.8 RL=100k	0–7.8 RL=10k
A_{VD} V/mV	15–200	15–200	20–500	15–280	7.5–40
CMRR dB	70–90	70–76	70–88	70–88	65–82
I_{CC}	0.5–3.3mA	1.4–2.8mA	10–40µA	150–400µA	1–2.2mA
B_1 MHz	0.7–1	3	0.1	0.7	2.3
SR V/µS	0.5	13	0.04	0.6	4.5
en nV/Hz @ 1kHz	22*	18	70	38	30
in pA$\sqrt{Hz}$ @ 1kHz	0.55*	0.01	0.013*	0.013*	0.013*

1) bipolar: µA741 / MC1458 / LM324
2) biFET: TL080-series
* typical value (not specified)
** TLC25X types are also specified at 1V

890039 - 11

Table. 1. Comparison between bipolar, BIFET and LinCMOS operational amplifiers.



DEVICE FEATURES Typical at $V_{DD} = 5\ V$, $T_A = 25°C$				
	BIAS-SELECT MODE			UNIT
	HIGH	MEDIUM	LOW	
P_D	3375	525	50	µW
SR	3.6	0.4	0.03	V/µs
V_n	25	32	68	nV/√Hz
B_1	1.7	0.5	0.09	MHz
A_{VD}	23	170	480	V/mV

Parameter Bias	Low	Medium	High
Offset voltage drift (Typ)*	0.1µV/month	0.1µV/month	0.1µV/month
Offset temp coef (Typ)	0.7µV/°C	2µV/°C	5µV/°C

890039 - 13

Fig. 2. The voltage at the bias-select pin of a LinCMOS operational amplifier influences the characteristics of the device.

Operational amplifiers

The operational amplifier is the most popular of all LinCMOS circuits. In fact, the first devices available in the new technology were the now well-known opamps Type TLC251/271 (single), TLC252/272 (dual), and TLC254/274 (quad). These are intended as replacements for the standard Types 741/3140, MC1458/CA3240, and LM324.

A close examination of the TLC271 reveals that the opamp, apart from the usual inputs, output and supply connections, has a so-called bias-select pin. The voltage at this pin determines the current drawn by the device—see Fig. 2. In the low-bias mode (bias-select pin connected to the +ve supply voltage), the current is only (typically) 10 µA. The price to be paid for this low current is a poor slew rate of only 0.04 V/µs and a unity-gain bandwidth of a mere 100 kHz.

The speed is determined largely by the device's internal capacitances. When the supply current is small, the charge and discharge currents through these capacitances will assume a larger importance, whence the lack of speed. However, there are a number of applications in which the low slew rate is of no importance

In the medium-bias mode, the current is about 15 times higher, but the slew rate and unity-gain bandwidth are correspondingly better: 0.6 V/µs and 0.7 MHz respectively). These values are comparable to those of, for instance, a standard 741. Note however that the latter draws a current of 1.7 mA.

In the high-bias mode, the current rises to 1 mA, but for that you get a very fast opamp with a slew rate of 4.5 V/µs and a

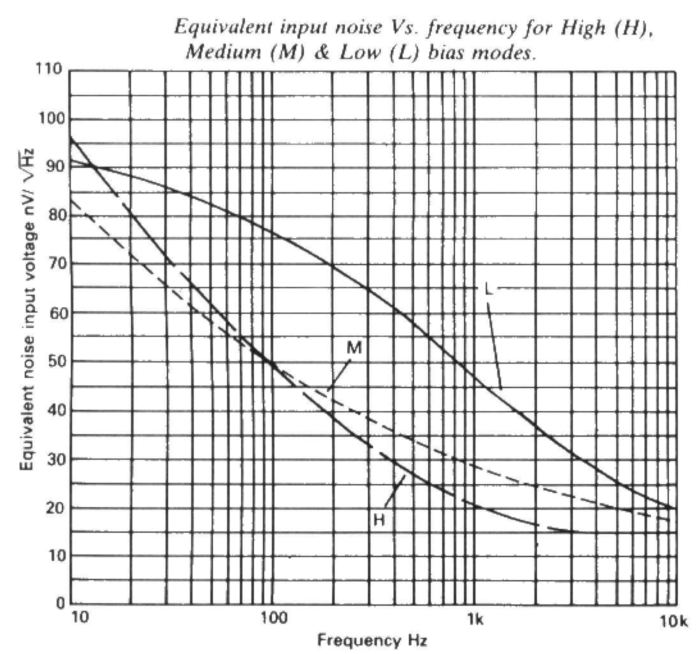


Fig. 3. Equivalent input noise voltage vs frequency for high (H), medium (M) and low (L) bias modes.

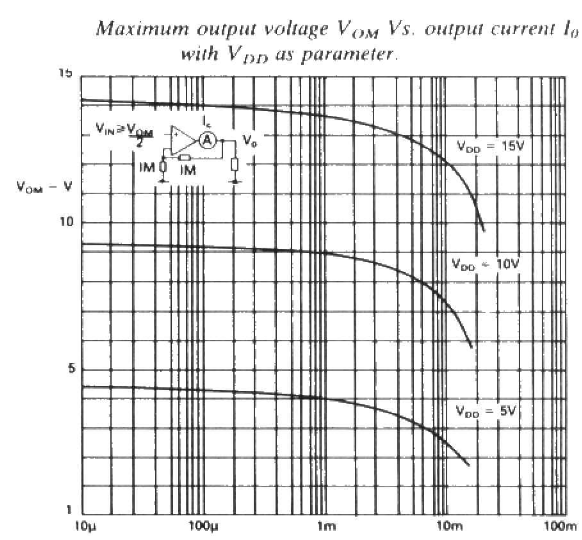


Fig. 4. Maximum output voltage V_{OM} vs output current I_O at various supply voltages.

unity-gain bandwidth of 2.3 MHz.

The dual versions TLC252/272 and the quad Type TLC254/274 are not provided with bias-select pins (for which they have no space in any case). In these versions, the bias mode is permanently set internally. The type number indicates which bias mode the device provides. For instance, a TLC27L2 is a low-bias type; a TLC272 is a high-bias version; and the TLC272M2 is a medium-bias type.

Power supply and load

As mentioned briefly already, to obtain maximum dynamic range, LinCMOS opamps are optimized for single-rail operation from supplies not exceeding 16 V.

The output voltage vs output current characteristic for loads connected to earth is given in Fig. 4. An open-circuit output, or one with the load connected to earth, can be pulled to within a few millivolts of 0 V. The output can only be pulled to the +ve supply level if the load is connected to the +ve rail or an external pull-up resistor is added. Such a resistor has, however, the disadvantage of resulting in a relatively large power consumption at low output voltages. Also, the open-loop amplification drops sharply when the output voltage gets close to the +ve supply voltage. This is caused by N5—see Fig. 1—switching off.

When relatively heavy loads are used, it should be noted that the sinking rate may exceed the sourcing rate; in other words, that the output current is greater than can be provided. If, therefore, large output currents are required without additional components, it is recommended that the load is connected to the +ve supply rail as shown in Fig. 5b.

Frequency compensation

In low-power applications, the current will be determined largely by the resistances in the feedback loop and by the load. The value of these resistance will, therefore, be normally quite high. As far as DC signals are concerned, that presents no problems. When AC signals are involved, however, more account must be taken of input and other stray capacitances (C_{stray} in Fig. 6) than in conventional opamp circuits. To obtain a sufficiently wide bandwidth, it may in some cases be necessary to use a compensating capacitor as shown in Fig. 6 to reduce the feedback at high frequencies.

Comparators

A number of comparators available in LinCMOS technology are shown in Table 2. The TLC372 and TLC393 are pin-compatible replacement of, for instance, double comparator Type LM393. Quad comparator Type LM339 may be replaced by the TLC339 or TLC374.

As with opamps, the current consumption of LinCMOS comparators is substantially lower than that of bipolar equivalents, while the input current is very low (typically 5 pA). There is not much difference in the input offsett voltages.

The outputs of most comparators are of the open-drain type, enabling logic functions to be produced by interlinking them. Normally, a pull-up resistor will also be required, but not with the TLC3702 and TLC3704, since these have totem pole outputs.

Unlike some opamps, comparators do not offer a choice of three bias modes. The bias mode is inherent in the type. For instance, the TLC393 draws 22 µA (typical) compared with the 0.8 mA drawn by an LM393, but it is slightly slower (2.5 µs agains 1.3 µs). A TLC372 has a higher power consumption, but is much faster (650 µs)

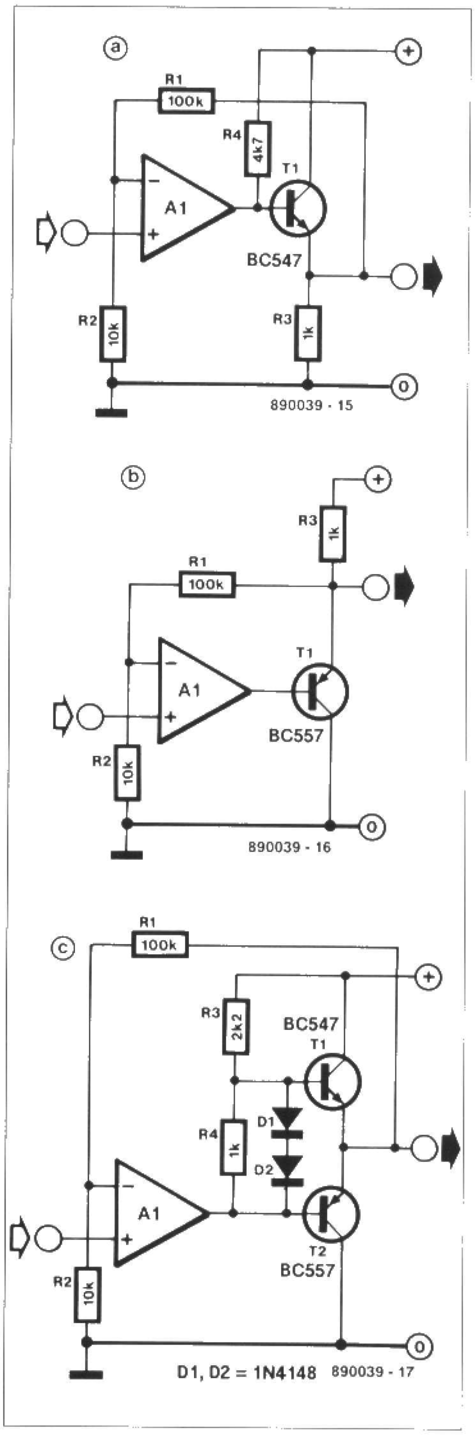


Fig. 5. Circuit adaptations for obtaining a larger output current in a load connected (a) to earth, (b) to the +ve supply rail, and (c) symmetrically.

COMPARATORS Table 2

	INPUT OFFSET VOLTAGE (mV)	INPUT OFFSET CURRENT (pA)	INPUT BIAS CURRENT AT 25°C (pA)	RESPONSE TIME (µS)	SUPPLY CURRENT (uA)	SUPPLY VOLTAGE RANGE (V) Min	Max	OUTPUT TYPE
Dual								
TLC372	12.0	1	5	0.65	750	3	16	Open Drain
TLC393	10	1	5	2.10	50	3	16	Open Drain
TLC3702	10	1	5	2.30	50	3	16	Totem Pole*
Quad								
TLC374	12.0	1	5	0.65	1000	3	16	Open Drain
TLC339	10	1	5	2.10	100	3	16	Open Drain
TLC3704	10	1	5	2.30	100	3	16	Totem Pole*

**Totem Pole Outputs are HCMOS and TTL compatible*

890039 - 21

Table 2. Some LinCMOS comparators and their main parameters.

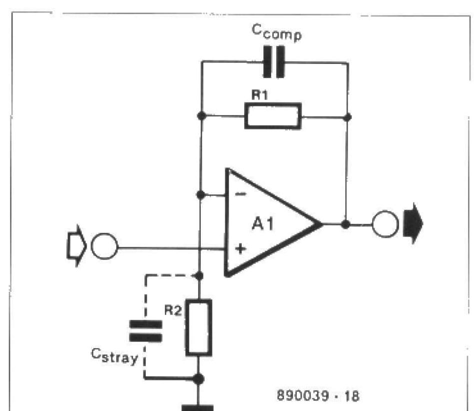


Fig. 6. Input and other stray capacitances may be countered by a compensating capacitor.

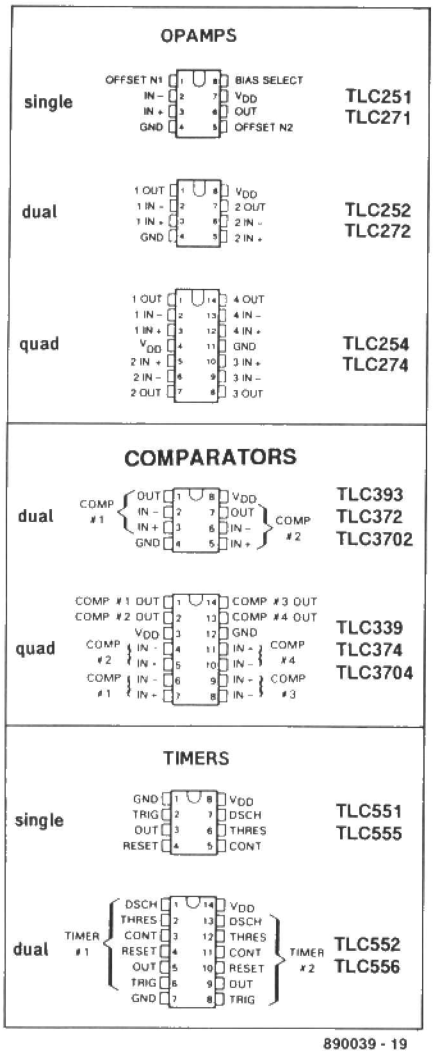


Fig. 7. Pinouts of the various LinCMOS circuits discussed in this article.

TIMERS — Table 3

	SUPPLY CURRENT (µA)	POWER DISSIPATION (mW)	SUPPLY RANGE (V) Min	Max	MAX FREQUENCY (MHZ)	MAX TIMING PERIOD	MAX TIMING ERROR	OUTPUT CURRENT (mA)
Single								
TLC551	350	1	1	18	2.1	Hours	3%	+10/–100
TLC555	350	1	2(3)*	18	2.1	Hours	3%	+10/–100
Dual								
TLC552	1000	2	1	18	2.1	Hours	3%	+ 10/–100
TLC556	1000	2	2(3)*	18	2.1	Hours	3%	+ 10/–100

**Indicates for industrial Temp Range*

890039 - 22

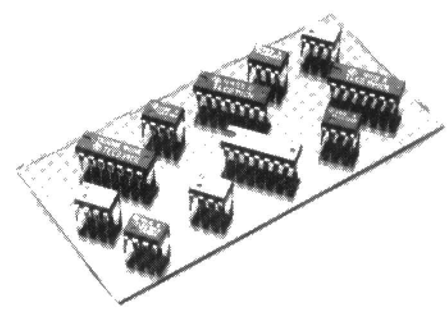
Table 3. A variety of LinCMOS timers and their main parameters.

It should be noted that Texas Instruments is not consistent in their recommendations on maximum supply voltage voltage. In some data sheets they mention 16 V and in others, 18 V. It is, perhaps, wise to be on the safe side and stick to 16 V.

Timers

As might be expected, there are equivalent LinCMOS ICs for the renowned 555 series of timers. In fact, there are four: the TLC555 (single) and TLC556 (dual) to replace the standard (LM/NE)555 and 556 for supply voltages from 2 V to 18 V and the TLC551 and TLC552 for operation from low voltages (down to 1 V).

Apart from the much reduced power consumption, the great benefit LinCMOS timers offer is the greatly extended frequency range. The maximum frequency is about ten times higher than that of a standard 555 (2.1 MHz against 200 kHz), because the saturation normal transistors have to cope with has no or negligible effect in the new technology. Even at relatively low frequencies (from 20 kHz to some hundreds of kHz), the TLC555 has a major advantage in that the frequency can be defined much more precisely by external components. Note that the frequency range is extended also at its lower end.



Since the input impedance and input leakage current are much smaller than in a bipolar 555, the RC networks that are connected to these timers can have a much higher value. This makes realization of very long time delays (up to hours)possible.

Examiner for hard-disk computer memories

Dissatisfaction with the available equipment led Servo Computer Services to develop its own system for examining the disks and disk-packs in computers.

The 'Disk Pack and Cartridge Inspector' can use either an existing disk drive or its own disk-drive spindle to test away from the drive. Highly accurate spindles are needed for this as the flatness of the disk surface needs to be within $\pm 0.5\times10^{-6}$ metre (2×10^{-5} inch) accuracy. In addition, such measurement is made on most disk without physical contact, so an optical method is used.

Hard disks often contain very valuable company information and the surface quality is very important. The Servo Inspector uses a variety of light sources as well as mirrors and lenses to carry this out.

The equipment has been designed to be used quickly and easily, to be reliable and rugged enough to be carried around by a service engineer. It fits easily into a compact case, suitable for air transport.

Servo Computer Services • 2–10 Sheffield Road • Dronfield • SHEFFIELD S18 6GG • Telephone (0246) 410375.

Versatile computer keyboard

A robust, easy-to-use keyboard, designed as a teaching aid for pupils and teachers with no technical computer knowledge, is available from Star Microterminals.

The 'Concept Keyboard' will interface with most computers in common use and has already been used extensively in schools for work on a wide range of subjects.

It uses a touch-sensitive membrane that is sensitive over the entire matrix area. Each touch cell produces an 8-bit output code and individual cells can be grouped to form large response areas. An EPROM option allows selective non-response areas.

Outside the educational field, the Concept keyboard has applications in word processing, banking, information management, robotics, stock control, etc.

Star Microterminals Ltd • Moorside Road • Winnal Industrial Estate • WINCHESTER SO23 7RX • Telephone (0962) 843332.

FLOPPY DISK MONITOR

M. Noteris

It often happens that PC users are left completely unaware of what is actually happening to the floppy disk inserted in the machine. Is the machine reading, attempting to read, or writing, and if so, to which track? This simple monitor circuit for IBM PCs provides the answers by making the control signals of the disk drives visible.

The drive select LED on a floppy disk drive does just what it is supposed to do: indicate drive activity. Many PC users have, therefore, no idea whether the floppy disk they have just inserted is read from or written to. Clearly, this is an unacceptable situation in this day and age of data security and a few bits on a disk determining access to files that represent many hours of work. While the present circuit can not restore data on a corrupted floppy disk, it helps to prevent the most serious of mishaps because you witness how they come about!

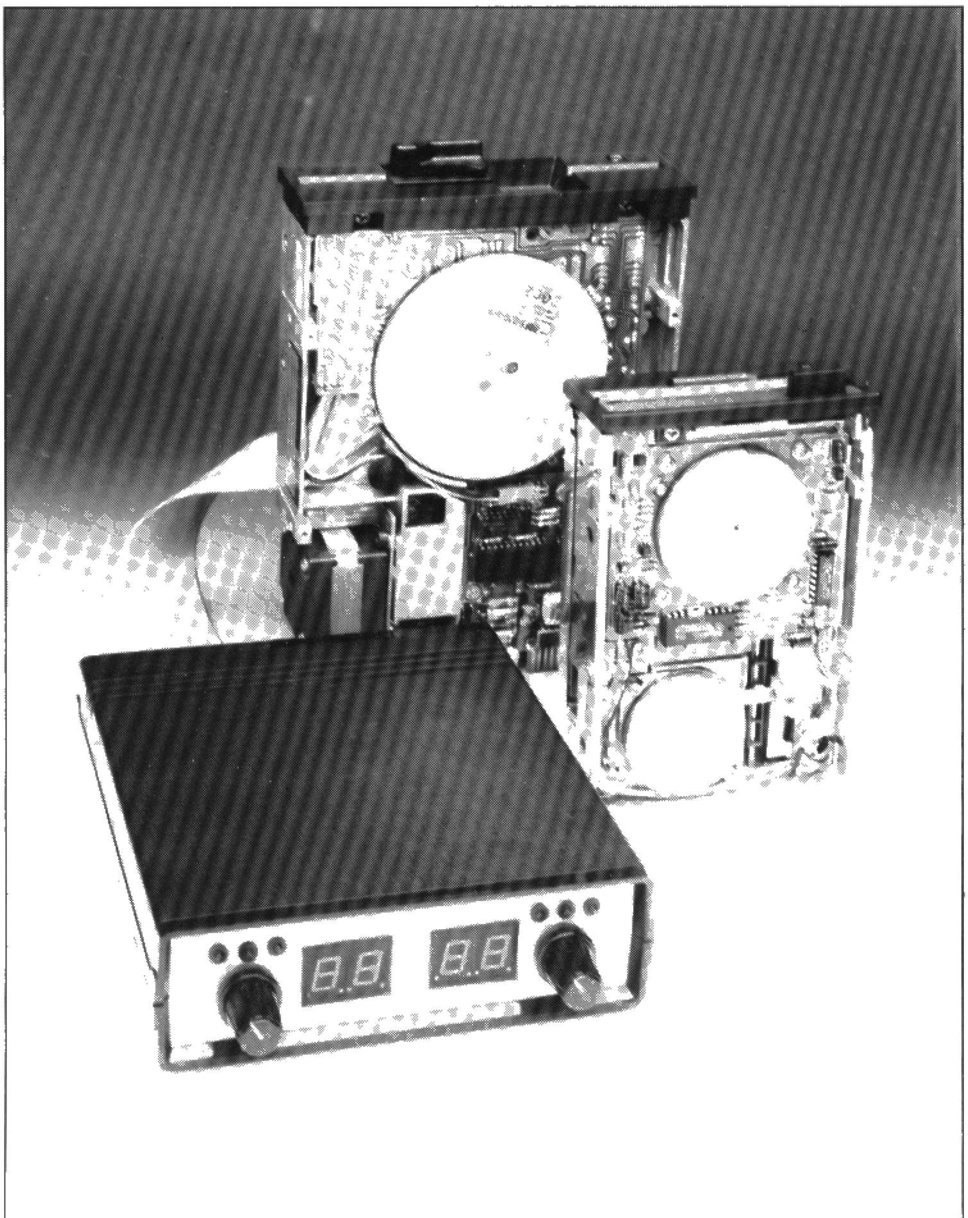
The principle

The floppy disk monitor works on the simple principle of visually indicating the status of the various control signals used for the floppy disk drives in a PC. Practically all user manuals supplied with PCs give a disk-drive wiring diagram that indicates the signals **Drive Select** (DS0 to DS2, and, in some cases, DS3), **Read Data**, **Write Enable**, **Step**, **Direction** and **Track 0**.

The movement of the head in the disk drive is fairly simple to monitor by clocking a counter with the Step pulses, driving the up/down input of the same counter with the Direction signal, and driving its reset input with the Track 0 signal. The visual indication function is assumed by a Type 4543 IC that decodes BCD data supplied by a counter Type 4510. The 4543 is capable of supplying the required 20 mA segment current for a Type 7760 LED display, of which two show the current track number.

Since the maximum number of tracks supported by the floppy disk monitor is 80 (0–79), two counter/display circuits are cascaded by driving the CARRY IN input of the decade driver with the CARRY OUT signal of the unit driver.

Signals Read Data and Write Enable are visualized with the aid of the decimal points on the LED displays. These indications are referred to as DPR (*decimal point read*), and DPW (*decimal point write*) in this article.



The circuit

The circuit diagram shown in Fig. 1 may cause some readers to wonder why two

- monitors all floppy disk drives available for PC/XT, PC/AT and compatible PCs: 5¼-inch, 3½-inch, single/double sided, double or quadruple density
- static display of head position (current track number)
- read and/or write indication for selected drive
- read indicator shows data flow resulting from pulses induced in the head by the magnetic carrier
- monitors two floppy disk drives simultaneously

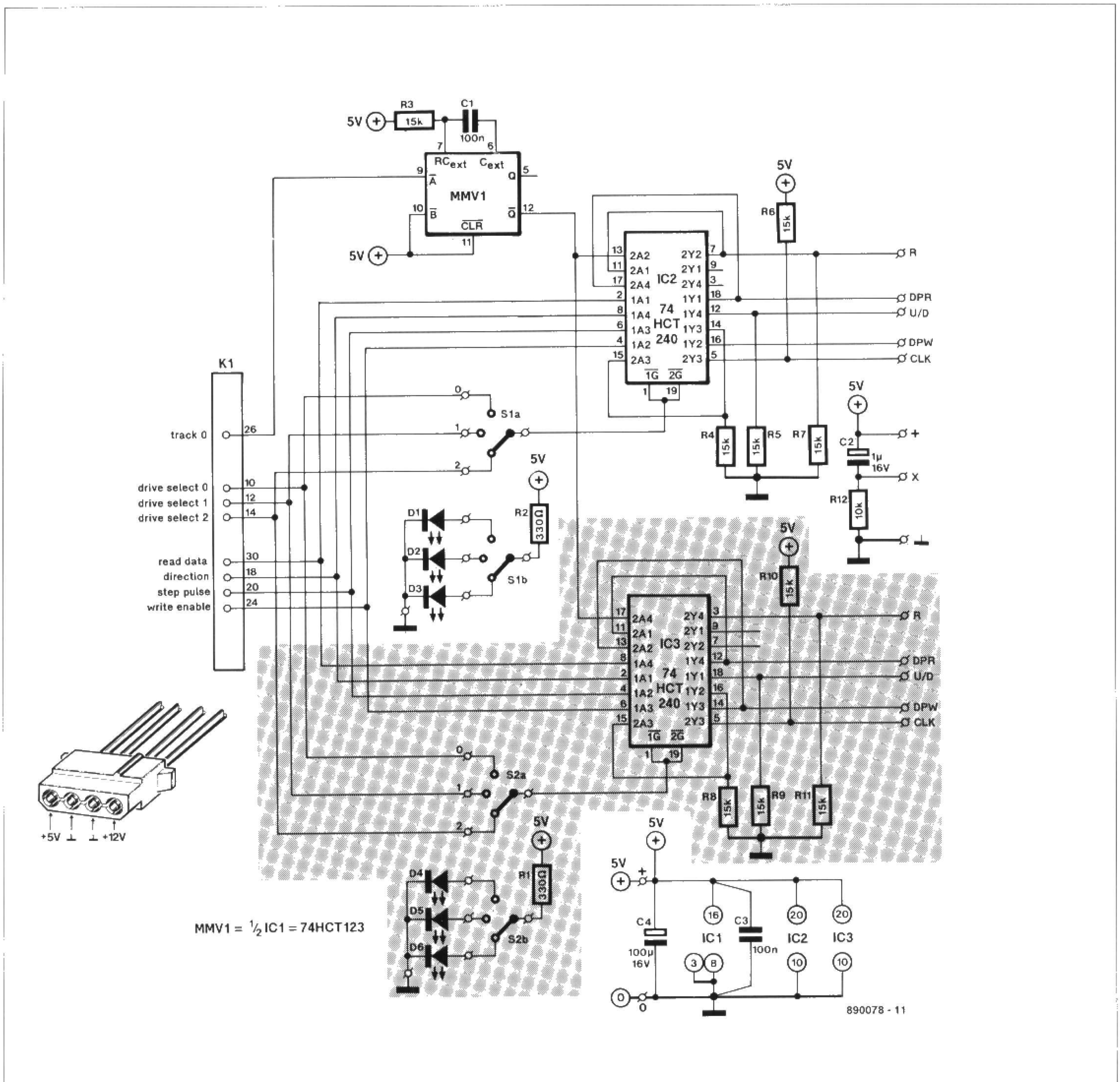


Fig. 1. Circuit diagram of the driver section of the floppy disk monitor. Shown inset is the pinning of the cable socket for powering disk drives.

Type 74HCT240 bus buffers are used. The six signals used for controlling all disk drives in a PC are carried in parallel between the disk controller board and the disk drives, so that the signals for the actuated drive must be selected before they can be directed to the associated indicator circuit. This directing of control signals is accomplished with the aid of the drive select lines, DS0, DS1 and DS2, which enable the bus buffers depending on the position of switches S_1 and S_2.

Since all control signals involved are active low, they are inverted by the 74HCT240s to enable driving the display units. A number of bus buffer inputs are connected to outputs to make sure they are properly terminated.

With S_1 in the position indicated in the circuit diagram, signal Drive Select 2 enables bus buffer IC_2 with a low level at inputs $\overline{1G}$ and $\overline{2G}$. Pull-up and pull-down resistors are fitted on the output lines of the bus buffers since these are switched to high-impedance when the device is disabled via the $\overline{1G}$ and $\overline{2G}$ inputs. Output lines R (reset) and U/D (up/down) are pulled low, and CLK (clock) is pulled high.

The Track 0 signal guarantees that the displays are always correctly reset to zero, which is useful when, for one reason or another, the counter loses track of the step pulses. Monostable multivibrator MMV_1 shapes the Track 0 signal supplied by a slotted optocoupler in the disk drive. The arm on which the head is mounted interrupts a light beam when it is in the extreme outer position with the head(s) over track 0 of the disk. The monostable lengthens the track 0 signal to about 1.5 ms to make sure that the display counter is properly reset even with drives that use the fastest step rate, 3 ms. Edge-triggering is used to cope with tolerance on the track 0 detection circuit and associated mechanical parts. In certain disk drives, the signal is still active even when the head is half way between tracks 0 and 1.

R-C network R_{12}-C_2 resets the counter circuits to zero at power-on by supplying a brief 'high' pulse via line X.

More about drive selection

Although the cable from the disk controller board has four drive select wires, DS0–DS3, the practical number of floppy disk

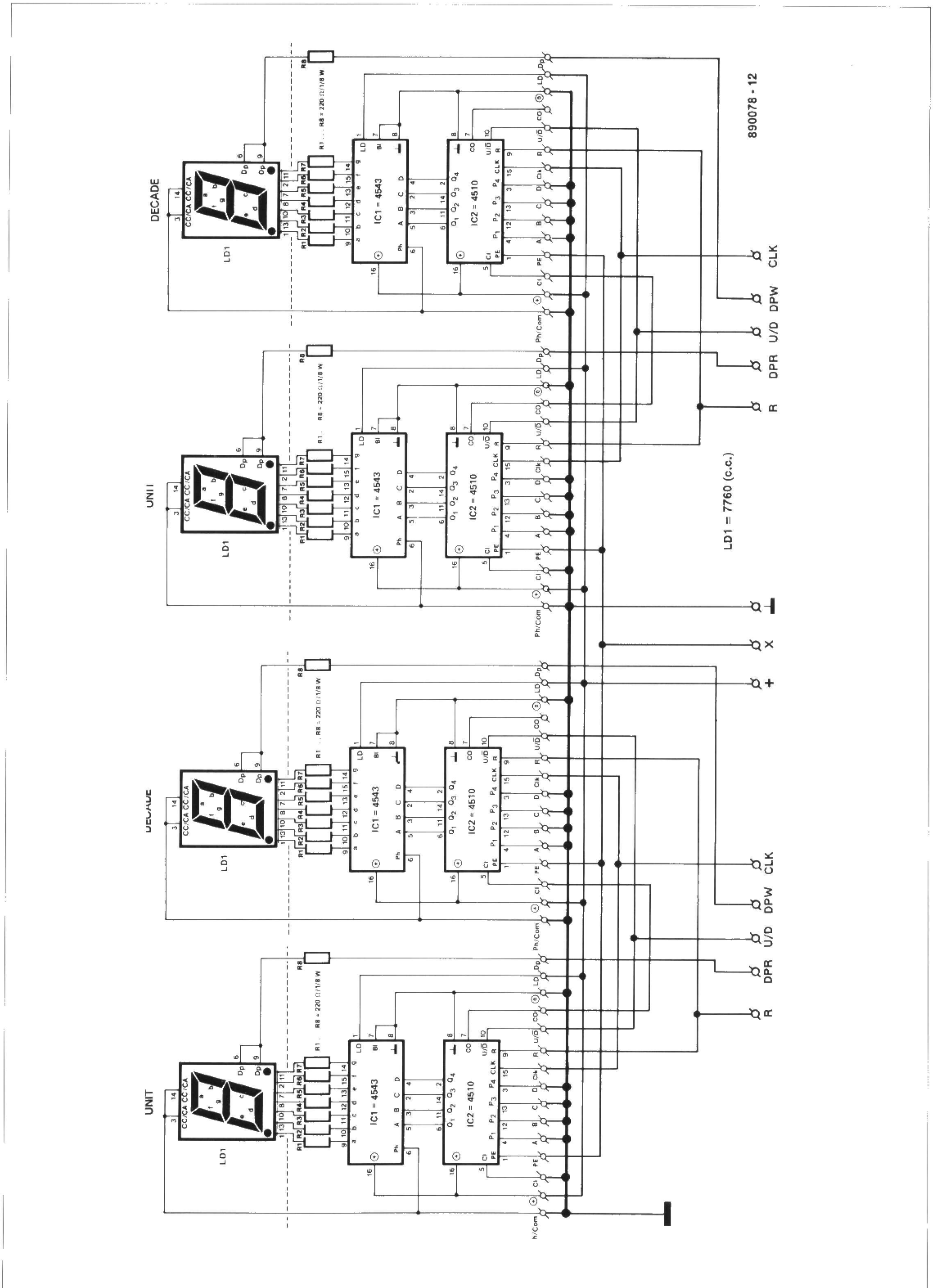


Fig. 2. Circuit diagram of the counter/display module.

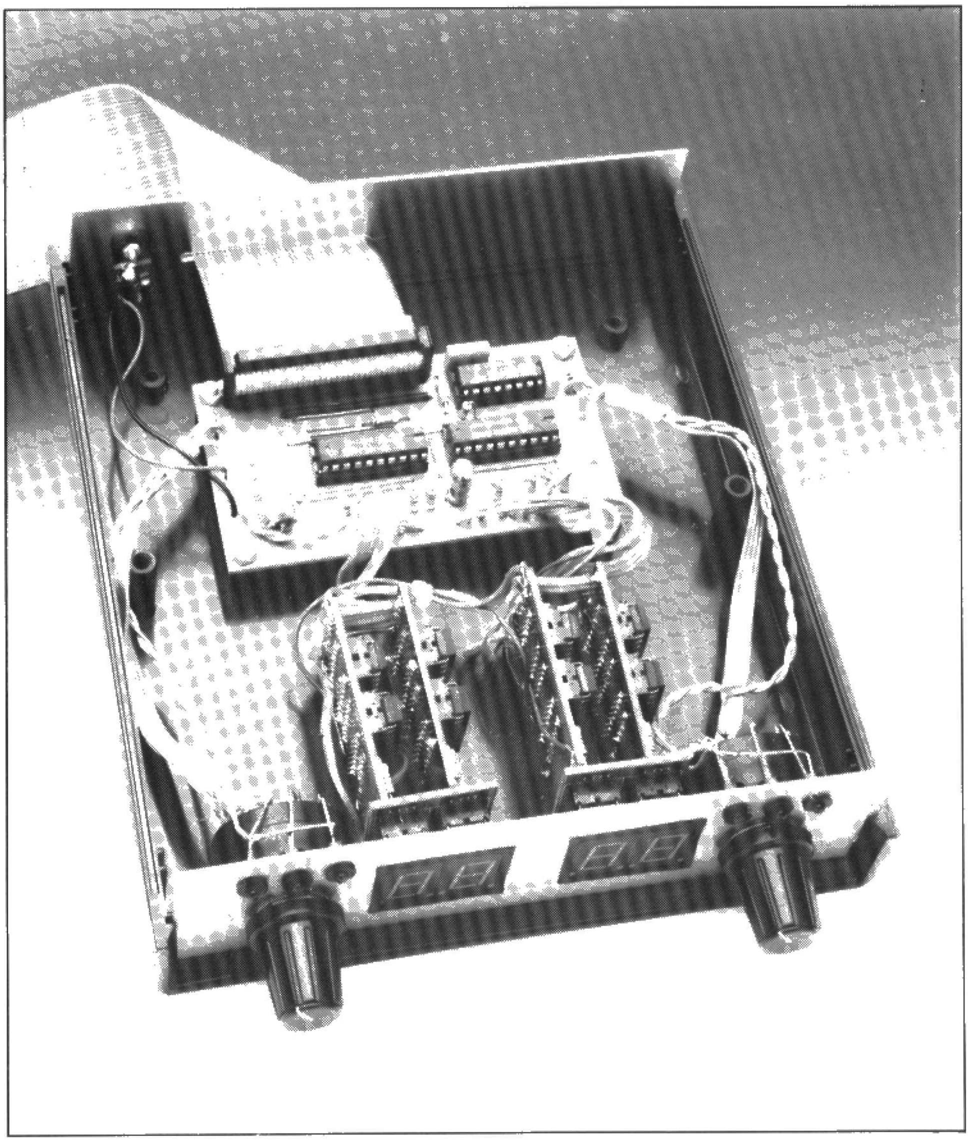
drives supported in IBM PCs and compatible machines is usually limited to two. This is because each floppy disk drive requires two signals, one to control the motor, and one to control the actual selection of the drive. Thus, the motor in drive A is energized under the control of a low level on line DS0, while the drive proper is enabled under the control of a low level on line DS2.

It should be noted that the above functions of DS0 and DS2 are the other way around on some PC compatibles of Far Eastern make. The floppy disk monitor solves a potential problem arising from this oddity by virtue of rotary switches S_1 and S_2.

Another noteworthy point is that signal Write Enable is fed to the display unit. The use of Write Data would appear more logical at first. The background to the use of Write Enable is that some disk controller boards, for instance, those of Western Digital, generate clock pulses on the Write Data line except when actually writing to the disk drive. This clock pulse stream can not be used by the display circuits, and would cause these to light the WRITE indication (DRW) continuously.

Counter and display module

The counter/display circuit is based on an earlier design published in Ref. 1. Figure 2 gives the circuit diagram. The module is composed of four identical combinations of a synchronous BCD counter Type CD4510, a latching BCD-to-7 segment display driver Type CD4543, and a common-cathode LED display Type 7760. Cascading is achieved by connecting



Parts list

DRIVER BOARD

Resistors (±5%):
R_1;R_2 = 330Ω
R_3–R_{11} = 15k
R_{12} = 10k

Capacitors:
C_1;C_3 = 100n
C_2 = 1µ0; 16 V; radial
C_4 = 100µ; 16 V

Semiconductors:
D_1–D_6 = LED; red; 3 mm
IC_1 = 74HCT123
IC_2;IC_3 = 74HCT240

Miscellaneous:
S_1;S_2 = two-pole, three-way rotary switch.
K_1 = 34-way pin header.
PCB Type 890078 (see Readers Services page).

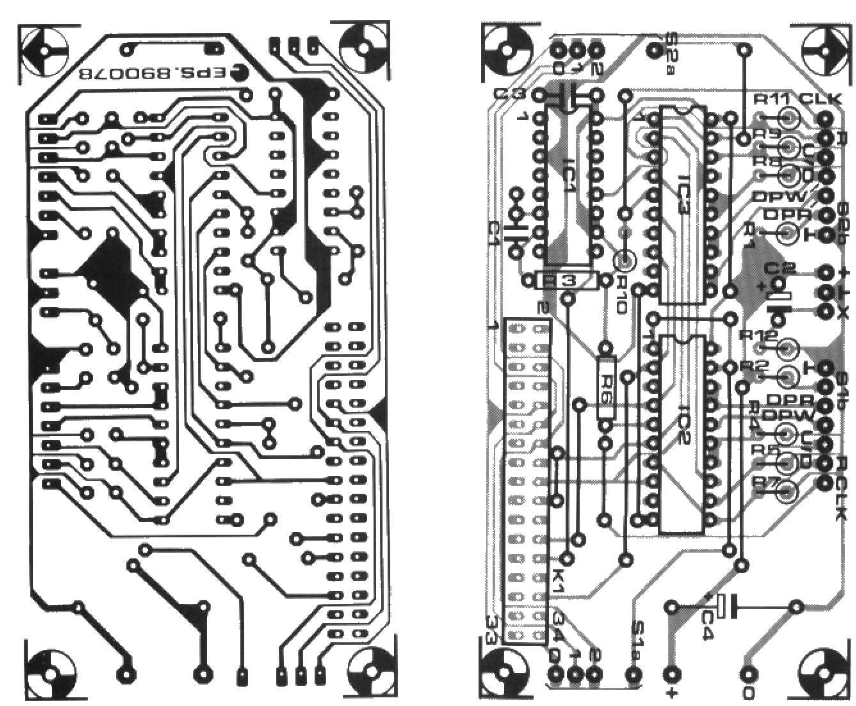


Fig. 3. Track layout and component mounting plan of the driver board.

Parts list

COUNTER/DISPLAY BOARD

Resistors (±5%):
R1–R8 = 220Ω

Semiconductors:
IC1 = 4543
IC2 = 4510
LD1 = 7760

Miscellaneous:
PCB Type 85019 (see Readers Services page; 4 off required for two display units).

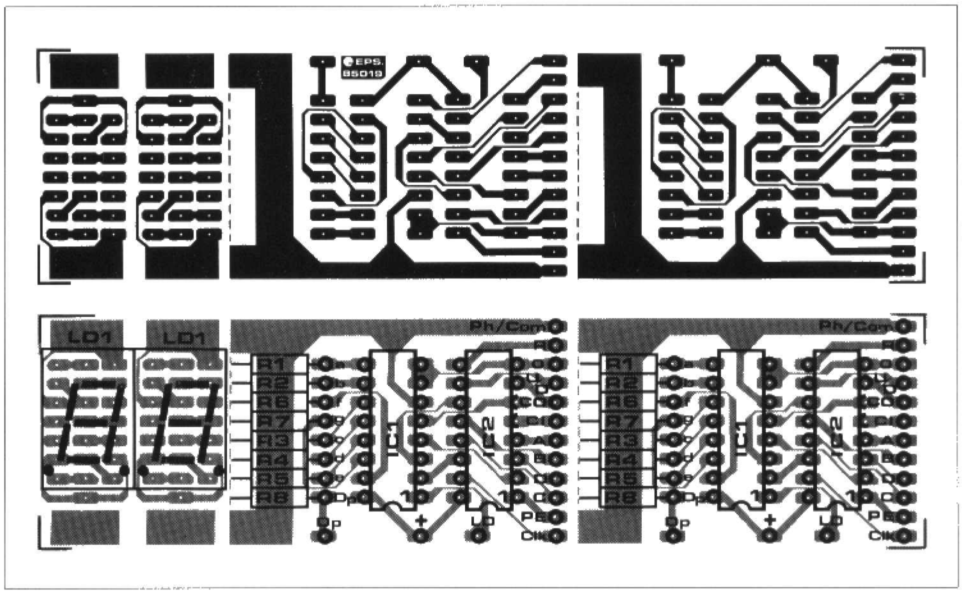


Fig. 4. Track layout and component mounting plan of the counter/display board.

the CARRY OUT (CO) output of each of the two units counters to the CARRY IN (CI) input of the associated decade counter. The functions of the U/D (up/down), and R (reset), Ph, BI, PE and LD are covered in Ref. 1.

Construction

The driver circuit of the floppy disk monitor is built on the printed-circuit board shown in Fig. 3, and the counter/display module on that shown in Fig. 4. Neither board should present any difficulty in populating.

Start the construction of the driver board by fitting the 12 insulated wire links. Continue with the resistors, of which most are fitted upright, the capacitors, the soldering terminals, IC sockets, and, finally, the 34-way pin header, in that order.

Each display board accommodates two displays and two driver circuits. The ready-made printed-circuit board must, therefore, be cut into three along the two dotted lines printed at the component side.

The fitting of the parts is carried out as usual. The interconnection between the two control boards to the display circuit requires further detailing, however. Taking one pair of displays as an example, the construction is started by populating the display board, and then the associated control circuit. Resistors R1 through R8 are mounted between the control board and the display board, and give the complete assembly the required rigidity.

Proceed with connecting paired points Ph/COM, R, 0, U/D, PE, Clk, + and LD. The CO output of the units display driver is connected to the CI input of the decade display driver, as shown in the circuit diagram of Fig. 2 (note the mirrored position of the displays in this drawing).

The completed counter/display units are connected to the driver circuit via the 6 signal lines and the 2 supply lines. In the standard version of the floppy disk monitor, there are two counter/display units and one driver unit.

The remaining connections are those for the LEDs and the rotary switches. Install the wiring as shown in the circuit diagram.

Power supply

The circuit is conveniently powered from the 5 V rail provided by the computer's power supply. The prototype of the floppy disk monitor is a stand-alone unit that is powered via a small socket as used on portable cassette recorders. The PC is fitted with a similar socket, and the two units are interconnected by a 30 cm long 2-wire supply cable.

If the circuit is installed permanently in the PC, the ground and +5 V connections may be made at appropriate points on the motherboard. Another, more practical, solution is shown in Fig. 7. A cable should be made to enable the supply voltage to be taken from one of the disk drives.

Cables

The floppy disk monitor is connected to the disk controller board via a home-made flat ribbon cable. This cable is crucial to the operation of the circuit and is, therefore, drawn in Fig. 5.

The job is almost done if the right parts are to hand: 50 cm or so of flat-ribbon cable, two 34-way IDC (*insulation displacement*) sockets, and one 34-way IDC header. The sockets and the header may be types with or without a strain relief clip. The two sockets are mounted at the cable ends, and the IDC header at about 15 cm from one of the sockets. Do not twist the cable in between the header and

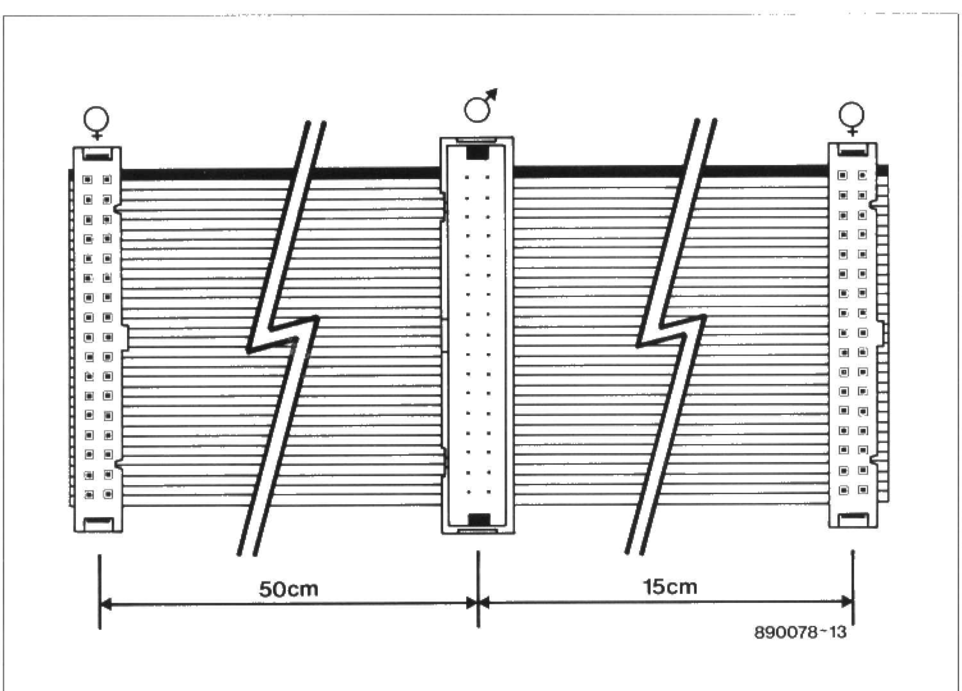


Fig. 5. The home-made 34-way flat-ribbon cable.

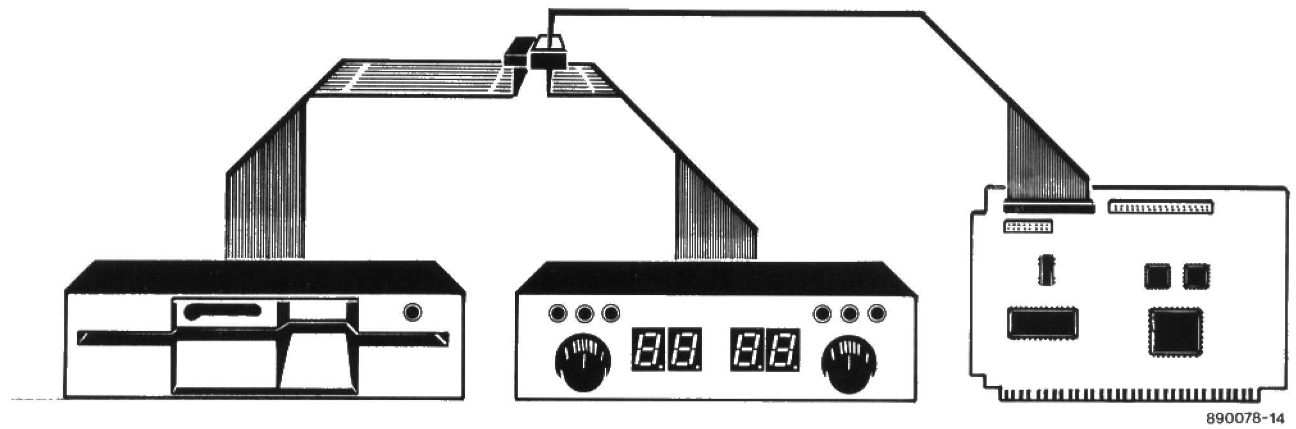


Fig. 6. Interconnection between the floppy disk drive, the monitor and the disk controller card in the PC.

the sockets: use the coloured wire in the cable to mark pin 1 of the connectors.

The existing cable between the floppy disk drives and the controller board must be disconnected at the side of the controller board. Figure 6 illustrates how the sockets at the ends of the previously described cable are connected to the disk controller board and the floppy disk monitor. The free end of the cable to the floppy disk drive is connected to the header on the home-made flat cable.

The connection as detailed is not affected by the number of floppy disk drives monitored with the present circuit. All activity on one, two or even three floppy disk drives may be watched closely from now on.

Modifications

As already noted, the basic version of the circuit is intended for monitoring the control signals of two floppy disk drives. It is, however, possible to realize a version for a single drive. In that case, only one universal counter module is used, while one of the two bus buffers on the driver board may be omitted.

A three-drive version of the monitor simply requires a third universal counter module, and, in addition, the shaded part of the circuit in Fig. 1. It is possible to mount the additional 74HCT240 on top of IC_2 or IC_3, soldering pins 2, 4, 6, 8, 10, 17 and 20 to the IC below, and bending the remaining pins away from the IC body for wiring as indicated in Fig. 1.

Unfortunately, IBM PCs and compatibles do not normally allow the use of more than two floppy disk drives. There are ways to overcome this limitation, but these fall outside the scope of this article. In an IBM environment, therefore, the floppy disk monitor can not keep an eye on more than two drives.

For computers that do support more than two floppy disk drives, the monitor circuit can be modified as required by using rotary switches with the corresponding number of positions.

The floppy disk monitor is not suitable for use with hard disks because these have a much higher number of tracks and heads.

In practice, the floppy disk monitor is a simple, yet effective aid for the PC user. It obviates, for instance, the use of a software utility to find out on which track a program or file is started, or how it is arranged on the disk.

Reference:

1. *Versatile counter circuit. Elektor Electronics* March 1985, p. 54–57.

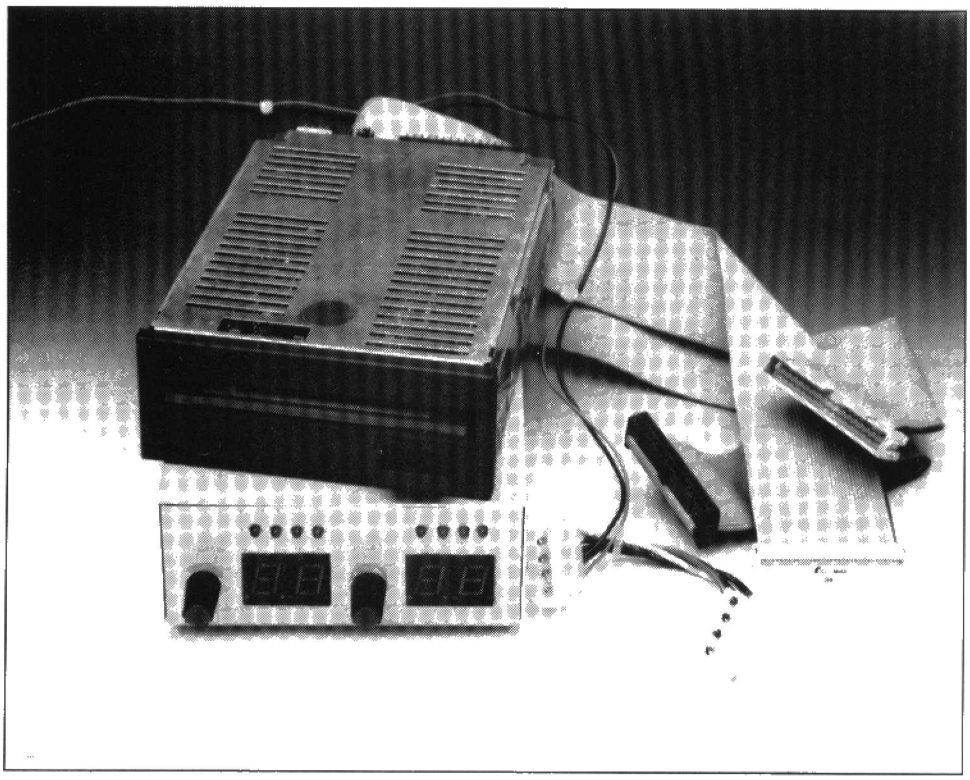


Fig. 7. The circuit is also perfect for use with 3½-inch disk drives.

PRACTICAL FILTER DESIGN - PART 7

by H. Baggott

After last month's discussion on Butterworth filters, we turn our attention in this seventh part of the series to a network that does not have such steep skirts, but makes up for that by its excellent pulse behaviour and very smooth transition characteristic: the Bessel filter.

The major advantage of Bessel sections is their phase behaviour, which is more linear than that of any other type of filter – see Fig. 38. If the Bessel amplitude characteristic is projected on a linear scale, it is a straight descending line. It is only because of its usual projection on a logarithmic scale that the characteristic looks like a typical filter skirt. The transfer characteristic of a Bessel filter is, therefore, moderate. A roll-off of $6n$ dB per octave is not attainable: the curve is particularly poor around the cut-off frequency and this is not dependent on n (n is the order of the filter).

Bessel tables

The tables giving the data for the computation of Bessel filters are compiled similarly to those for Butterworth sections, described in detail last month, and they should therefore not present any problems.

Again, the component values are given for a cut-off frequency of 1 Hz. Table 6 gives the pole locations of Bessel filters from the 2nd to the 10th order, while Tables 7, 8 and 9 give the component values for a passive section under different operating conditions.

Bessel characteristics

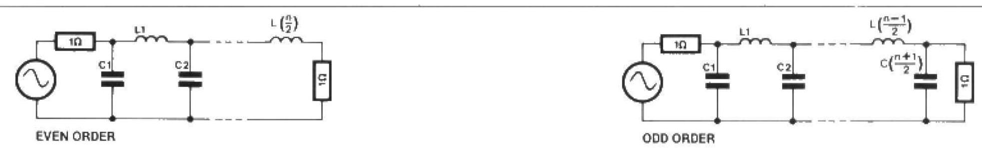
The characteristics in Fig. 37–39 show the plus and minus points of a Bessel filter. For instance, the roll-off is noticeably less steep than that of a Butterworth filter. Also, the knee is virtually the same for all orders of the filter. One of the positive qualities of the Bessel filter is seen in Fig. 38. The delay vs frequency characteristics

CORRECTION

Tables 2 and 3 in last month's instalment contained some errors. The drawings inset in those tables included the descriptions 'even orde' and 'oneven orde'. These should have read 'even order' and 'odd order' respectively.

n	real part $-\alpha$	imaginary part $\pm\beta$
2	1.103	0.6368
3	1.0509	1.0025
	1.327	
4	0.9877	1.2476
	1.3596	0.4071
5	0.9606	1.4756
	1.3851	0.7201
	1.5069	
6	0.9318	1.664
	1.3836	0.9727
	1.5735	0.3213
7	0.9104	1.8375
	1.3797	1.1923
	1.613	0.5896
	1.6853	
8	0.8955	2,0044
	1.3787	1.3926
	1.6419	0.8253
	1.7627	0.2737
9	0.8788	2,1509
	1.3683	1.5685
	1.6532	1.0319
	1.8081	0.5126
	1.8575	
10	0.8688	2,3002
	1.365	1.7395
	1.6667	1.2248
	1.8477	0.7297
	1.9335	0.2423

Table 6. Pole locations of Bessel filters.



n	C1	L1	C2	L2	C3	L3	C4	L4	C5	L5
2	0.0916	0.3418								
3	0.0537	0.1545	0.3507							
4	0.03715	0.107	0.1721	0.3566						
5	0.02774	0.08072	0.128	0.1768	0.3594					
6	0.02172	0.06369	0.1017	0.1359	0.1771	0.3604				
7	0.0176	0.05187	0.08354	0.1117	0.1383	0.1759	0.3606			
8	0.01463	0.04327	0.07017	0.09447	0.1162	0.1384	0.1744	0.3606		
9	0.01241	0.03681	0.06	0.0813	0.1004	0.1179	0.1375	0.1729	0.3605	
10	0.0107	0.0318	0.05204	0.07089	0.08798	0.1033	0.1181	0.1363	0.1716	0.3603
	L1	C1	L2	C2	L3	C3	L4	C4	L5	C5

Table 7. Normalized component values for passive low-pass filters with identical input and output impedances.

n	L1	C1	L2	C2	L3	C3	L4	C4	L5	C5
2	0.2167	0.07224								
3	0.2329	0.1341	0.04657							
4	0.2389	0.1557	0.09751	0.03365						
5	0.2407	0.1628	0.1199	0.07526	0.02575					
6	0.2407	0.1644	0.1293	0.09664	0.06024	0.02048				
7	0.2401	0.1638	0.1328	0.1075	0.08007	0.04954	0.01677			
8	0.2394	0.1626	0.1336	0.1127	0.0914	0.06769	0.04163	0.01405		
9	0.2388	0.1612	0.1331	0.1149	0.09775	0.07899	0.05816	0.03562	0.012	
10	0.2383	0.1599	0.1321	0.1155	0.1011	0.08596	0.06911	0.05064	0.03091	0.01039

Table 8. Normalized component values for passive low-pass sections with negligible source impedance.

n	C1	C2	C1	C2	C3
2	0.1443	0.1082			
3			0.2265	0.1572	0.04039
4	0.1611	0.006207			
	0.117	0.1047			
5	0.1657	0.04934			
			0.1607	0.1387	0.04926
6	0.1708	0.04076			
	0.115	0.07695			
	0.1011	0.09708			
7	0.1751	0.03444			
	0.1154	0.06607			
			0.1358	0.124	0.04818
8	0.1776	0.02956			
	0.1155	0.05714			
	0.09693	0.07737			
	0.09029	0.08817			
9	0.181	0.02591			
	0.1163	0.05025			
	0.09626	0.06926			
			0.1204	0.1125	0.04538
10	1.1832	0.02287			
	0.1166	0.04444			
	0.09549	0.06201			
	0.08613	0.07452			
	0.08231	0.08104			

Table 9. Normalized component values for active filters with a single feedback path.

are highly linear up to the cut-off frequency (from about the 3rd order onwards). With higher orders, the delay remains linear for some time beyond the cut-off frequency. This linear behaviour is also found in the step response in Fig. 39: there is virtually no overshoot or sign of ringing.

Some examples

Example 1.

Third-order low- and high-pass Bessel filters are required for a loudspeaker system with a nominal impedance of 8Ω. The cut-off frequency is required to be 2500 Hz.

Solution.

It is assumed that the source impedance is negligibly small and that the required filter is a passive one. The computation of the low-pass section is simplicity itself. We take a standard passive low-pass filter and insert the component values for a 3rd-order section from Table 8—see Fig. 40a. Subsequently, those values are transferred to the real load impedance (8Ω) and the actual cut-off frequency (2500 Hz):

$$C' = C/(fR)$$

$$L' = LRf$$

The resulting section is shown in Fig. 40b.

Next, the high-pass filter. All capacitors in the low-pass filter are replaced by inductors and all inductors by capacitors. In Part 3 we have seen that normalized filter values of a high-pass section are found by 'inverting' the normalized values of a low-pass filter. For the present example, this has been done in Fig. 40c with the addition of a factor $4\pi^2$, which is necessary because the formulas $1/C$ and $1/L$ apply to normalized values for $\omega = 1$ rad/s, whereas in the tables in these articles the standard values are given for $f = 1$ Hz, whence the correction factor. The formulas thus become:

$$C_h = 1/(4\pi^2 L_l)$$

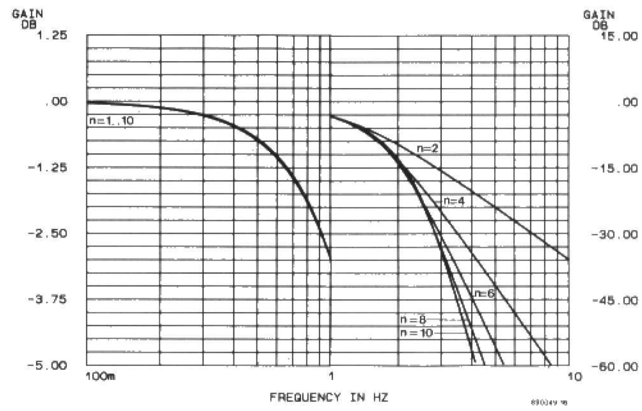


Fig. 37. Gain vs frequency characteristics of a Bessel filter.

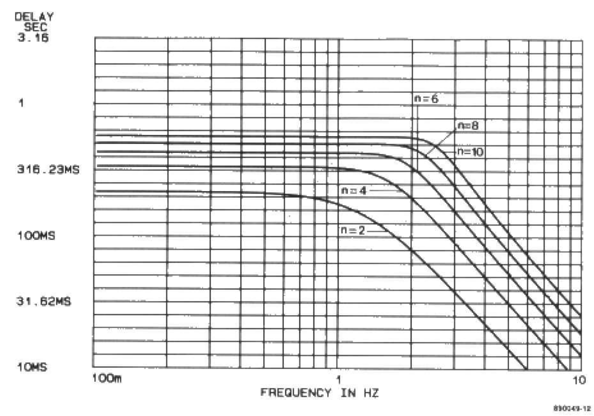


Fig. 38. Delay time vs frequency characteristics of a Bessel filter.

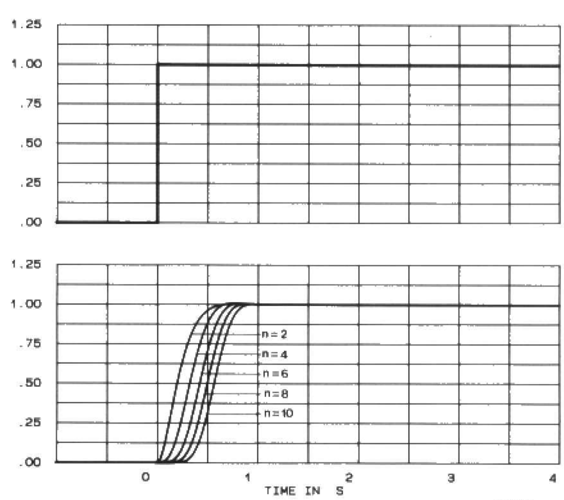
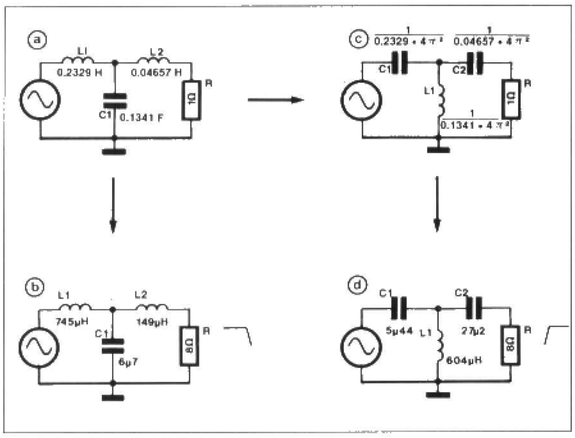


Fig. 39. Step response of a Bessel filter.



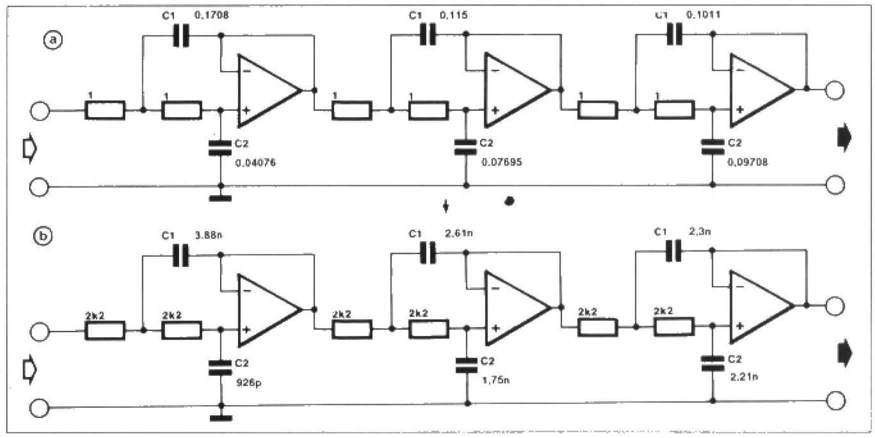


Fig. 40. Dimensioning a passive low- and high-pass filter: load impedance is 8Ω and cut-off frequency is 2500 Hz.

Fig. 41. An active Bessel filter of the sixth order; cut-off frequency is 20 kHz.

$L_h = 1/(4\pi^2 Cl)$

Since a calculator is required anyway, it is quite simple to include the factor ($4\pi^2 = 39.48$).

The computation of the resulting high-pass filter terminated into 8Ω and having a cut-off frequency of 2500 Hz is then carried out in the usual way—see Fig. 40d.

Example 2.
A low-pass Bessel filter is required with a cut-off frequency of 20 kHz. The time delay must remain constant up to not less than 30 kHz. The attenuation at 100 kHz must be at least 50 dB.

Solution.
First, the delay and attenuation frequencies must be converted to 1 Hz and this gives values of 30/20=1.5 and 100/20=5. Next, we ascertain which order of filter meets the requirements.

In Fig. 38 we see that a filter of about the sixth order is required to keep the delay constant up to around one-and-a-half times the cut-off frequency. It is then seen in Fig. 37 that a sixth order filter is required to give an attenuation of at least 50 dB at $5 \times f_c$.

A sixth-order filter is an even-order one, so it must be constructed from 2nd-order sections—see Table 9. The capacitor values for the three sections may be taken direct from Table 9. Next, choose a suitable value for the resistors. The actual capacitor values and the wanted cut-off frequency are then computed with the aid of the formulas used in the first example—see Fig. 41b.

Two calculations for the first section:

$$C_1 = 0.1708/(20{,}000 \times 2200) = 3.88 \times 10^{-9} = 3.88 \text{ nF.}$$

$$C_2 = 0.04076/(20{,}000 \times 2200) = 926 \times 10^{-12} = 926 \text{ pF.}$$

Next month: Chebishev filters.

EVENTS

IEE Meetings
- 3–6 July — Software engineering for telecommunication switching systems (Bournemouth).
- 7–9 July — History of electrical engineering (Swansea).
- 12–14 July — Optical fibre communication (Univ. College, London)
- 16–21 July — Transmission for telecommunications (Aston Univ).
- 18–20 July — Image processing and its applications (Univ. of Warwick)
- 19 July — Annual General Meeting.
- 24–27 July — Microwave (Sao Paulo, Brazil).
- 22–25 Aug — Antennas and propagation (Tokyo).

Readers are also advised that each year the IEE organizes a range of conferences, vacation schools, technical seminars and workshops on subjects within the fields of electrical, electronic and control engineering, and computing.

Further information on these, and many other, events from IEE • Savoy Place • LONDON WC2R 0BL Telephone 01-240 1871.

The INMARSAT **Conference on Mobile Satellite Communications** will be held at the QEII Centre, Westminster, London, from **17 to 19 July.** Further information from Blenheim Online Ltd • Blenheim House • Ash Hill Drive • PINNER HA5 2AE • Telephone 01-868 4466.

The **Light and Sound Show** will be held at Olympia 2, London, on **10–13 September**. Further details from the Professional Lighting and Sound Association • 1 West Ruislip Station • RUISLIP HA4 7DW • Telephone (0895) 634515.

The **Third International Symposium on IC Design and Manufacture** will be held on **14–15 September** in Singapore. Further details from the School of Electrical and Electronic Engineering • Nanyang Technological Institute • Nanyang Avenue • SINGAPORE • Telephone Singapore 2263.

The **European Microwave** exhibition will be held at the Wembley Conference and Exhibition Centre on **4–7 September**. Details from Microwave Exhibitions and Publishers • Telephone (0892) 44027.

The **Ineltc** exhibition will be held at Basle on **5–8 September**. Details from Sekretariat Bufa • Postfach CH–4021 Basle • Switzerland.

The **Internepcon Semiconductor** exhibition will take place in Hong Kong on **6–9 September**. Details from Reed Exhibitions, London • Telephone 01-940 3777.

The **Sonimag** exhibition will be held at Barcelona on **11–17 September**. Details from ADG Exhibitions, Chesham • Telephone (0494) 729406

UNIVERSAL MIDI KEYBOARD INTERFACE

Final Part

D. Doepfer

Table 3 shows a practical example of how the EPROM contents can be adapted to the available keyboard, in this case a 72-key type with a range from *F* to *E*. As discussed last month, programming the EPROM is simply a matter of entering the actual key numbers in ascending order, starting with the lowest key. Note that the most-significant nibble of each programmed byte is 0, 2, 4, 6 or 8. The practical connection of this keyboard is shown in Fig. 6b in last month's instalment. Since there are 72 keys, 5 decoder boards are required to cover the 6 octaves.

Programming one's own EPROM is, fortunately, not required in most cases, since the EPROM supplied through the Readers Services (ESS575) provides the data required for a 96-key *C*-to-*B* keyboard (8 octaves). Most keyboards are smaller and will not require a reprogrammed EPROM if their contacts go to the correct inputs on the decoder boards.

Table 3.

EPROM contents for 72-key keyboard. Range: F to E (6 octaves)

	A8	A7	A6																	
n	0	0	0	0000	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
o	0	0	0	0010	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
t	0	0	0	0020	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
	0	0	0	0030	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
u	0	0	1	0040	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
s	0	0	1	0050	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
e	0	0	1	0060	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
d	0	0	1	0070	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF

DOWN-TRANSPOSITION (A7=1; A8=0)

*C = *middle C* on the first additional line under the treble stave.

	0	1	0	0080	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					C		D		E	F		G		A		B	C		D	
	0	1	0	0090	FF	00	01	02	03	04	05	06	07	08	09	0A	0B	0C	0D	0E
					E	F		G		A		B	C		D		E	F		G
L	0	1	0	00A0	0F	20	21	22	23	24	25	26	27	28	29	2A	2B	2C	2D	2E
						A		B	C		D		E	F		G		A		B
O	0	1	0	00B0	2F	40	41	42	43	44	45	46	47	48	49	4A	4B	4C	4D	4E
					C		D		E	F		G		A		B	*C		D	
W	0	1	1	00C0	4F	60	61	62	63	64	65	66	67	68	69	6A	6B	6C	6D	6E
					E	F		G		A		B	C		D		E	F		G
	0	1	1	00D0	6F	80	81	82	83	84	85	86	87	FF	FF	FF	FF	FF	FF	FF
						A		B	C		D		E	F		G		A		B
	0	1	1	00E0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					C		D		E	F		G		A		B	C		D	
	0	1	1	00F0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					E	F		G		A		B	C		D		E	F		G

UP-TRANSPOSITION (A7=0; A8=1)

*C = *middle C* on the first additional line under the treble stave.

	1	0	0	0100	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					C		D		E	F		G		A		B	C		D	
	1	0	0	0110	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					E	F		G		A		B	C		D		E	F		G
H	1	0	0	0120	FF	FF	FF	FF	FF	FF	FF	FF	FF	00	01	02	03	04	05	06
						A		B	C		D		E	F		G		A		B
I	1	0	0	0130	07	08	09	0A	0B	0C	0D	0E	0F	20	21	22	23	24	25	26
					C		D		E	F		G		A		B	*C		D	
G	1	0	1	0140	27	28	29	2A	2B	2C	2D	2E	2F	40	41	42	43	44	45	46
					E	F		G		A		B	C		D		E	F		G
H	1	0	1	0150	47	48	49	4A	4B	4C	4D	4E	4F	60	61	62	63	64	65	66
						A		B	C		D		E	F		G		A		B
	1	0	1	0160	67	68	69	6A	6B	6C	6D	6E	6F	80	81	82	83	84	85	86
					C		D		E	F		G		A		B	C		D	
	1	0	1	0170	87	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					E	F		G		A		B	C		D		E	F		G

NORMAL (A7=A8=1)

*C = *middle C* on the first additional line under the treble stave.

	1	1	0	0180	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					C		D		E	F		G		A		B	C		D	
N	1	1	0	0190	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	00	01	02
					E	F		G		A		B	C		D		E	F		G
O	1	1	0	01A0	03	04	05	06	07	08	09	0A	0B	0C	0D	0E	0F	20	21	22
						A		B	C		D		E	F		G		A		B
R	1	1	0	01B0	23	24	25	26	27	28	29	2A	2B	2C	2D	2E	2F	40	41	42
					C		D		E	F		G		A		B	*C		D	
M	1	1	1	01C0	43	44	45	46	47	48	49	4A	4B	4C	4D	4E	4F	60	61	62
					E	F		G		A		B	C		D		E	F		G
A	1	1	1	01D0	63	64	65	66	67	68	69	6A	6B	6C	6D	6E	6F	80	81	82
						A		B	C		D		E	F		G		A		B
L	1	1	1	01E0	83	84	85	86	87	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					C		D		E	F		G		A		B	C		D	
	1	1	1	01F0	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF	FF
					E	F		G		A		B	C		D		E	F		G

Construction

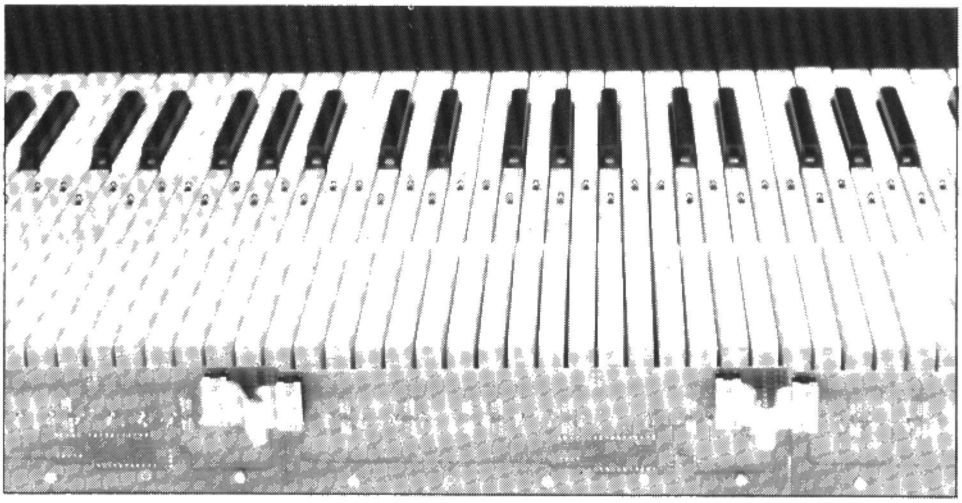
The construction of a MIDI keyboard entails mechanical as well as electronic work. It is important to note that the larger part of the descriptions that follow are based on the construction of a 72-key keyboard with wooden keys and spiral spring contacts. This is but an example, however, and many other configurations and constructions are possible. Keyboards with and without keys may be obtained from music stores.

Main controller board

Figure 7 shows the main controller board. Start the construction by fitting the six wire links on it (excluding J_1 and J_2) because a number of these are covered later by IC sockets. Use insulated wire to prevent short-circuits with the socket pins.

Voltage regulator IC_5 must be fitted on a small heat-sink if the input voltage applied to socket K_2 is higher than about 10 V (C_9 is omitted in that case). Wire links J_1 and J_2 connect address lines A9 and A10 to ground if the used part of the EPROM is located in the lower addressable range (this is so in most cases). Diodes D_3 and D_4 are mounted upright (note the orientation). Resistors R_{11}–R_{14} are contained in a single-in-line (SIL) array of which the 4 other internal devices are not used. The resistor array may be replaced by 4 vertically mounted resistors whose top connections are commoned by a horizontal wire which goes to the hole next to pin 1 of the EPROM.

Transposition switch S_1 must be not be installed direct on to the board because its pinning does not correspond to that of the solder terminals (the centre contact must be wired to point c). The MIDI connector is a 5-way DIN type for mounting at the edge of the printed circuit board.



Decoder boards

The number of decoder boards required depends on the keyboard type. One decoder board, of which the design is shown in Fig. 8, scans up to 16 keys, and is configured to address a particular group of keys with the aid of a jumper, J_1, which goes to point A, B, C, D, E, or F as shown in Fig. 6. Each of the cascaded decoder boards has a different jumper installed. Jumper A is always used for the decoder board connected to the lowest key group. If the total number of keys on the keyboard is not a multiple of 16, as, for instance, in the case of a 72-key type, only the *extreme right-hand* decoder board may be cut in two. The track design of the decoder board allows this to be done fairly easily to the right of contact S_8.

Connections

The controller board is connected to one of the decoder boards with the aid of a 16-way flat-ribbon cable. Both ends of this cable must be fitted with an IDC-type 16-pin DIP header. As shown in Figs. 6a and 6b, the cable connects socket K_1 on the controller board to socket K_3 on the nearest decoder board. The decoder boards are interconnected in ascending order with cables between sockets (or pin headers) K_2 and K_1. These are shown in dotted lines on the component overlay, and must, therefore, be installed *at the track side of the board*. Wire-wrap sockets or pin headers should be used to make sure that the pins can be soldered properly to the tracks. Wire-wrap sockets mate with 16-way IDC

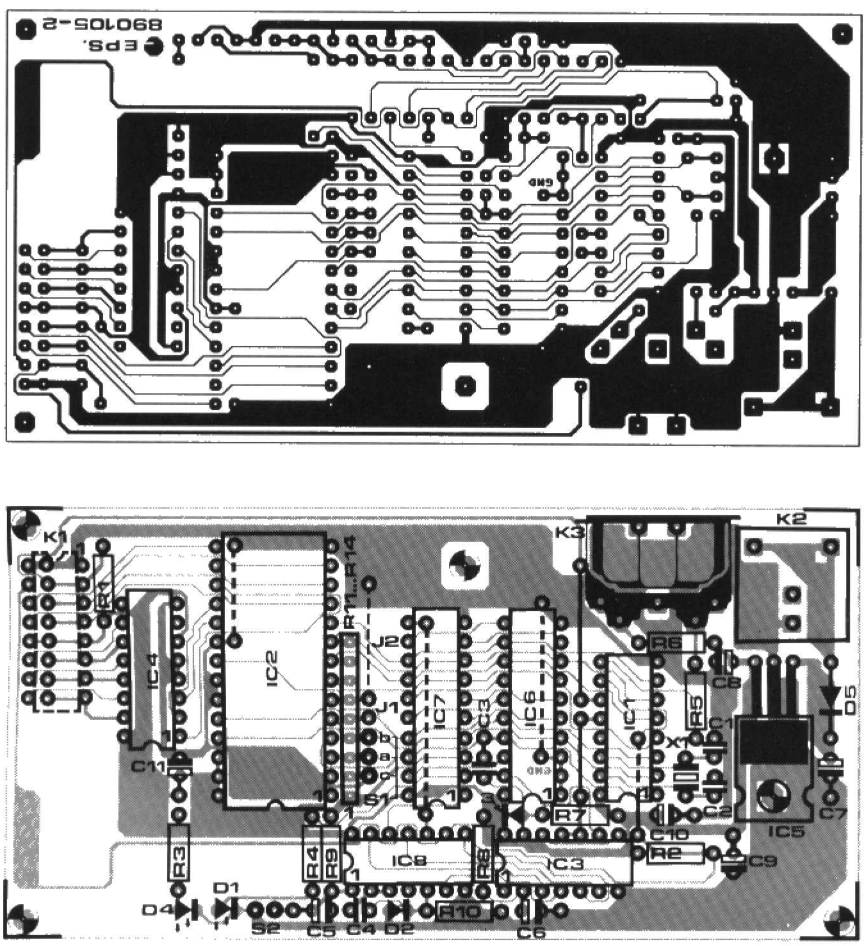


Fig. 7. Track layout and component overlay of the main controller board. The unregulated supply voltage is applied via K2.

Parts list

CONTROLLER BOARD

Resistors (±5%):
R1;R2 = 1K0
R3;R4 = 470R
R5;R6 = 220R
R7 = 10K
R8;R9;R10 = 100K
R11–R14 = 1K0 (8-way SIL array)

Capacitors:
C1;C2 = 22p
C3 = 100p
C4 = 10n
C5 = 1µ0; 16 V; tantalum
C6 = 10µ; 16 V
C7–C11 = 2µ2; 16 V; tantalum
C9 is not fitted when IC5 is mounted with a heat-sink.

Semiconductors:
D1;D4 = LED
D2;D3 = 1N4148
D5 = 1N4001
IC1 = E510⁺
IC2 = EPROM 2764 (ESS575; see Readers Services page)
IC3 = 74HC00 or 74HC132
IC4 = 74HC138
IC5 = 7805
IC6 = 74HC688
IC7 = 74HC273
IC8 = 74HC04 or 74HC14
Note: HCT equivalents may be used, but HC and HCT types must not be mixed in this circuit.

Miscellaneous:
X1 = quartz crystal 4 MHz.
S1 = miniature toggle switch with centre-off contact.
S2 = push-to-make button.
K1 = 16-way DIL socket or 2 8-way pin headers.
K2 = PCB mount socket for 3.5 mm supply plug.
K3 = PCB mount 5-way DIN socket.
Heat-sink for IC5.
PCB Type 890105-2 (see Readers Services page).

⁺Available from C-I Electronics, P.O. Box 22089, 6360 AB Nuth, Holland.

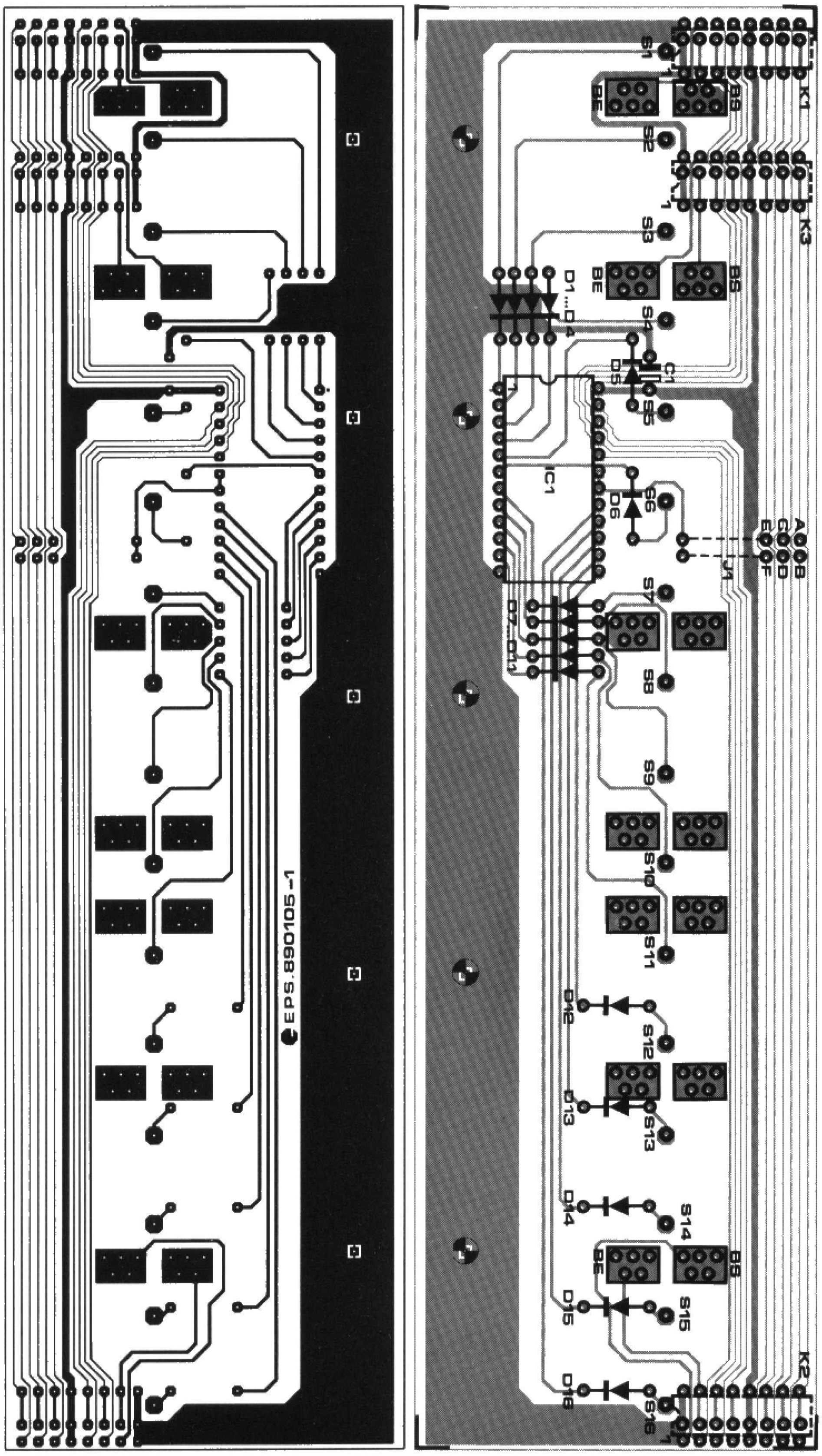


Fig. 8. Track layout and component mounting plan of a keyboard decoder. The BE and BS bus bars are fitted on to solder terminals, for which five possivle positions are provided. Sockets or pin headers K1 and K2 are mounted at the track side of the board.

Parts list

DECODER BOARD

Capacitors:
C_1 = 1µ0; 16 V; tantalum

Semiconductors:
D_1–D_{16} = 1N4148
IC_1 = 74HC154 or 74HCT154

Miscellaneous:
K_1;K_2;(K_3) = 16-way pin header or wire-wrap socket.
16-way flat-ribbon cable with IDC or DIP connectors.
Solder terminals as required.
PCB Type 890105-1 (see Readers Services page; quantity as required).

The prototype keyboard, a 6-octave type, was obtained through ELV France (see advert on p. 8/9).

DIP headers, and pin headers with IDC sockets. It is also possible to do without flat-cable connectors altogether and use individual wires. With 3, 4 or 5 decoder boards, and 16 wires per connection, however, this solution is rather time-consuming.

When a keyboard with fewer than 96 keys is used, and it is not desired to change the contents of the default EPROM (ESS575), study the configuration diagram of Figs. 6a, and find the decoder board and point S_1–S_{16} to which the lowest key contact on your keyboard is connected. Figure 6a shows examples of how keyboards with 72 keys and 54 keys are 'fitted' into the available range of 8 octaves. Again, note that the *standard* EPROM is used (ESS575), so that programming is not required. Non-connected contacts may be left open or tied to the BE bus to simulate a closed rest contact.

Mechanical work

Since there are many types of new, used and otherwise ready-made keyboards around, the mechanical construction is not standard as is the electronic construction.

The choice of the keyboard and the contact type is fairly difficult, and the best way to avoid problems is, of course, to purchase a keyboard with integral switchover keys. Unfortunately, such a device is probably hard to obtain, so that the keyboard and the contacts may have to be purchased separately. Gold-plated wire contacts are simply glued on to the decoder boards prior to soldering their contacts. Keyboards with wooden keys and spring-type contacts are probably the best you can get, but they require great care and precision in assembling successfully with the decoder boards.

Although the decoder boards (Fig. 8) have been designed to fit in line with the contacts and the keys themselves, they are also suitable for use with wooden keys

and separately installed contacts, as shown in the accompanying photographs. The BS and BE lines are constructed as bus bars that run along the full length of the cascaded decoder boards. The spiral springs touch the upper bar when the keys are pressed. The bars are made from silver-plated wire fitted on to solder terminals.

Each decoder board has holes to enable it to be secured on to the keyboard with screws and PCB spacers. The BE and BS bus bars or the contacts must not be used for keeping the boards in place. At least 3 of the 5 screw holes must be used: the two at the edges and the one in the centre of the board. The cross-sectional drawing of Fig. 9 gives a suggested construction of a keyboard with wooden keys and spiral spring contacts.

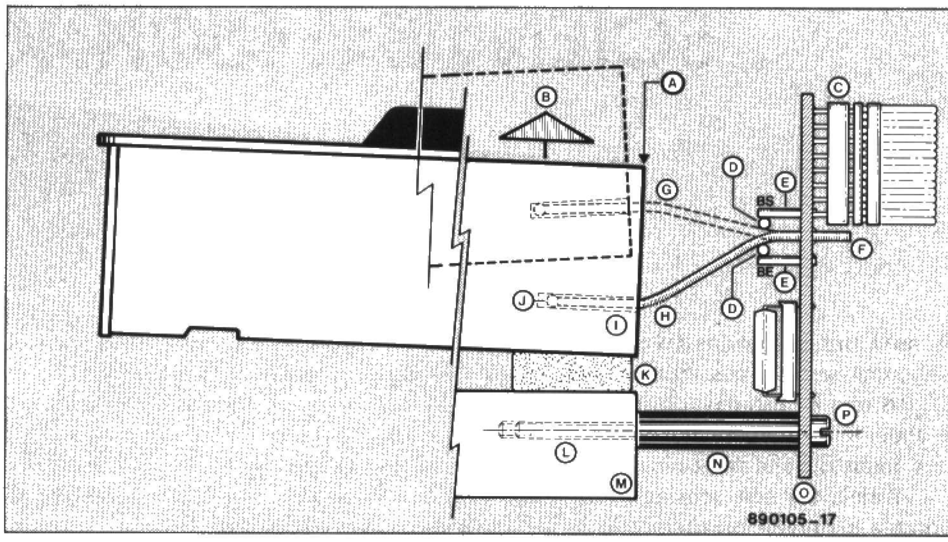


Fig. 9. Suggested construction using wooden keys and spiral springs. A: wooden key. B: key travel. C: wire-wrap socket. D: silver-plated bus bar. E: solder terminal. F: spring end soldered at track side of decoder board. G: spriral spring in actuated position. H: spiral spring in rest position. I: hole in wooden key. J: spring secured in key with super-glue. K: support strip for keys. L: hole drilled in keyboard base plate. M: keyboard base plate. N: PCB spacer. O: decoder board. P: metal screw.

Horizontal bus bars

To begin with, do not install the solder terminals for the BE and BS bus bars, or the contacts. Use an unpopulated decoder board to mark its position on to the rear of the keyboard. Wood screws may be used in some cases, and metal screws with PCB spacers in others, depending on the construction, size and material of the rear side and the base plate or frame of the keyboard.

The decoder boards allow 5 positions for the solder terminals that hold the bus bars. As will be recalled from Part 1 of this article, the keyboard processor measures the player's strike force as the time that lapses when the pole of the actuated key travels from the rest contact to the work contact. The key release time is established in a similar manner by measuring the work-to-rest time. Since all rest contacts are connected to bus bar BE, and all work contacts to bus bar BS, it will be evident that the 5 positions of the bars allow the constructor to gear the velocity characteristic very accurately to the player's preference (are you the constructor *and* the player? Good!).

While fitting the solder terminals and the bus bars, it is important to observe the distance between them, and the distance between the key contact (the spiral spring) and the BS bus bar. These distances must be uniform over the full length of the bars. A vernier slide gauge or a micrometer must be used to check this. In practice, it was found that a bar distance of 3 mm gave optimum results with the prototype keyboard shown in the photographs. Position errors of less than 0.5 mm were clearly noted by experienced players.

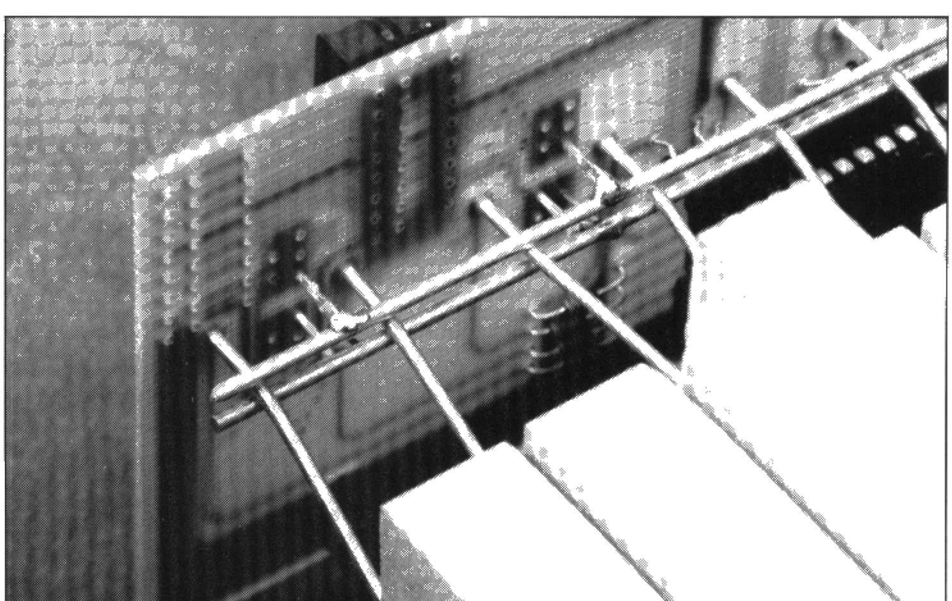
To find the optimum position of the bars on your particular keyboard, fit only 2 solder terminals on a single decoder board, that is provisionally secured on to the keyboard. Experiment with the bus distance until you are satisfied with the velocity response of the keyboard, then install the bars permanently and check for uniform spacing. The use of silvered wire is a must to prevent corrosion of the contacts. Silver-plated wire is usually sold in diameters from 1 to 2 mm.

The key pole should not reach the work contact, i.e, the BS line, until the key is fully down. This requires accurate positioning of the decoder boards, and is best achieved by making the screw holes slightly oval at the top with the aid of a small round file. Do not make the holes in the keyboard itself any larger.

The position of the BS bus bar with respect to the pole is fairly critical. If the pole reaches the work contact too early, adjacent keys will be actuated along with the wanted ones while playing rapidly. Also take into account the relatively large inertia of wooden keys, which take quite some time to return to the rest position. In general, if the work contact is not located at the very end of the pole travel, notes will blend (no NOTE OFF message) during rapid, but still *staccato*, playing. The problem in this case is common to all mechanical keyboards: the limit of the repetition rate is reached.

In conclusion, the work contact as well as the rest contact must be actuated neatly and reliably at the end of the pole travel, and at all times.

The drawing of Fig. 9 shows a suggested construction on the basis of wooden keys and spiral spring contacts. It is recommended first to glue the spring in a hole in the key, and then solder the other end to the decoder board. After the glue has hardened, the spring is pulled gently just before soldering it at the track side of the relevant decoder board. This is done to make sure that the spiral remains straight when the associated key is struck. Spring movement, however small, is unacceptable in the rest position. When movement is noted, heat the solder joint and carefully pull the spring a little to get more tension. Try out the feel of the keys by playing a few notes within the octave. Do not fit or adjust the remainder of the keys until the results of this test are satisfactory.



SCIENCE & TECHNOLOGY

The versatile cable that tells a tale

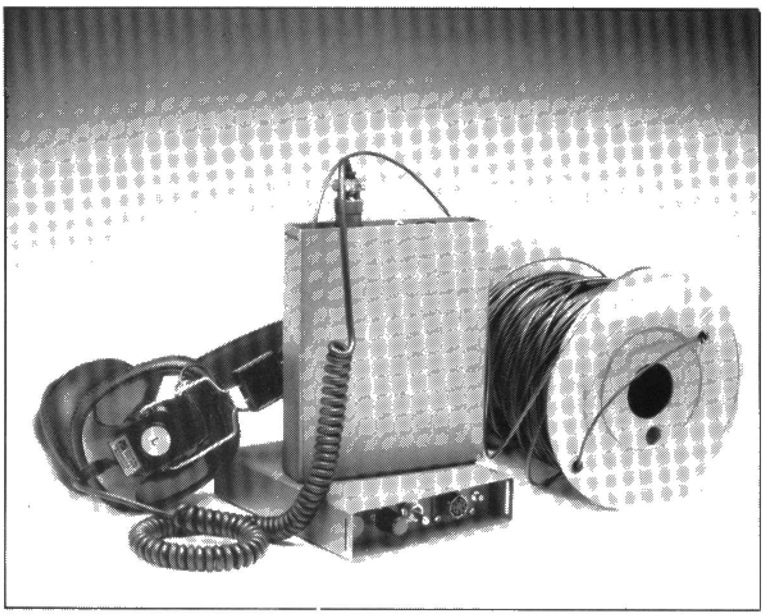
by Dennis Moralee

A new pressure-sensitive cable with important security possibilities, developed as part of a British naval research project, is about to open up a whole range of military, industrial and domestic applications.

Vibetek, the new sensing cable, incorporates a rugged piezoelectric polymer that allows it to respond directly to applied pressures, impacts, stress and strain, vibration and even sound. Compared to traditional piezoelectric sensing devices, using single crystals or microchip thin films that are limited to sensing at a single point location, Vibetek provides continuous sensing along cable runs of over several hundred meters. It also offers a much more sensitive response, is robust and competitively priced.

The new cable may be used simply to replace existing pressure-sensitive lines in, for example, traffic monitoring applications, where it can be buried just below a road surface in order to count the number of vehicles passing over it.

Its high sensitivity makes it possible for the cable to be put about 30 cm or more below the road surface, protecting it from physical damage while still providing accurate and reliable detection of passing vehicles. Its rugged construction will ensure that it survives for two years or more where conventional cables would last only a small part of this time.



Highest vibration frequencies

Moreover, it can detect not only the large applied pressures of road vehicles but the lighter loads as well. And because its response is highly linear and its electrical output precisely proportional to the applied pressure, different forms of road traffic can be distinguished and accurately weighed, proving the facilities of a dynamic weighbridge.

It is also sensitive over a wide range of frequencies, which in traffic-sensing applications can stretch from the near-static loads of slow-moving vehicles to the highest vibration frequencies transmitted through the road surface. Linked to an electronic signal processing instrument, the cable can provide information about a passing vehicle's weight, speed, load distribution, acoustic noise level, and even its engine condition. In suitable circumstances, it can also monitor the conversation of passing pedestrians.

The capabilities of the Vibetek cables are even more important in military and civilian security applications. For example, in one military situation, an army patrol, camping overnight in rough terrain, used the new cable as a trip wire along a lengthy security perimeter that included not only roads, farm tracks and fields, but also sections of running streams.

Through monitoring, the patrol could detect and track the movements of every type of intruder sent against it—ranging from a single infantryman to a whole platoon and wheeled and tracked vehicles.

Security functions

In everyday life, the security provided by Vibetek's capabilities is noteworthy. A length of cable attached to a perimeter fence around an industrial site, for instance, can pinpoint all attempts to tamper with it.

Buried in a floor or wall or even simply run under a carpet, it will detect all movements in a protected building, and may even pick up intruders' conversations, providing the perfect antidote to nimble-fingered thieves of artworks and security safes.

Raychem Ltd[1] of Swindon, the original developer of Vibetek, has set up a subsidiary company to develop the market both for the cable and complete electronic systems incorporating its use. Called FOCAS[2], the new firm is staffed mainly by former Raychem personnel. Raychem is a minority shareholder with the majority owned by security specialists, Cookson Group PLC[3].

FOCAS has already built up considerable interest in Vibetek among many British companies, particularly in the security field, and new commercial uses for the cable are expected to reach the market soon.

It is currently being supplied in two forms: an extremely sensitive premium grade called Vibetek 20, and a more rugged, cheaper version, Vibetek 5. Both use a piezoelectric layer of polyvinylidene fluoride (PVDF), formed coaxially between inner and outer electrodes.

In the case of Vibetek 20, these are made from a special alloy and silver respectively as part of a unique co-extrusion process. In Vibetek 5, however, the PVDF layers are conventional copper components surrounded by an extra rugged protective jacket. Both forms of cable allow easy connection through conventional cabling techniques, yet have very impressive technical characteristics.

References:

1. Raychem Ltd • Faraday Road • Dorcan • SWINDON SN3 5HH • Telephone (0793) 28171.
2. FOCAS • Faraday Road • Dorcan • SWINDON SN3 5HH • Telephone (0793) 28171.
3. Cookson Group PLC • Clements House • 14 Gresham Street • LONDON EC2V Telephone 01-606 4400.

TRACKING TESTER

This high-grade piece of test equipment, designed by ELV GmbH, can be used for tape speed calibration, wow, flutter and drift measurements on reel-to-reel tape recorders, cassette recorders and video recorders.

A close study of the tape tracking characteristics, together with signal-to-noise and frequency response measurements, is of undisputed importance for aligning audio and video tape recording equipment, and for quality assessment. Tracking errors are caused by irregular tape transport, and are manifest as frequency deviations in the played back signal.

At least two standards are available for tape tracking measurements. One is the DIN standard (*Deutsche Industrie Norm*), which is based on a test frequency of 3150 Hz. The other standard has been defined by the CCIR (*Comité Consultatif Internationale de Radio*), and is based on a test frequency of 3000 Hz. The tracking tester described here supports both standards by providing either test frequency at the flick of a switch.

Tests and procedures

A basic distinction is made between two test procedures:

Speed deviations and drift
This measurement requires a test cassette with a prerecorded reference signal of 3150 Hz (DIN) or 3000 Hz (CCIR). The accuracy of the frequency of this reference tone is adequate for all practical purposes, and enables measuring absolute tape speed deviations as well as drift. The latter is particularly noticable at the beginning and the end of the tape, when the reel motor(s) have to compensate relatively rapid changes in the load and torque. The operation of the tracking tester for this type of measurement is detailed below.

Wow and flutter
These tracking errors consist of slow frequency variations in the range from 0.3 to 6 Hz (*wow*), and as faster modulation effects up to 100 Hz, which results in a fluttering sound. The tracking tester has three ranges for wow and flutter measurements.

Depending on the selected mode and range, the tracking tester measures the relative deviation of the instantaneous tape speed (wow and flutter), or the relative deviation of the average tape speed (drift). The relative deviations are expressed as a percentage of the relevant standard values. As already mentioned, drift measurements require a test cassette.

Operation

The tracking tester is powered by a 9 VDC/200 mA mains adaptor connected to a socket on the rear panel of the enclosure.

The power on/off switch is found at the right on the front panel. A light-emitting diode (LED) above the switch indicates whether the instrument is in operation.

The required test standard, CCIR (3000 Hz) or DIN (3150 Hz), is selected with the aid of a single switch at the centre of the front panel, near the top cover of the enclosure. The signals to and from the tape recorder are connected to the tracking tester either separately via phono sockets or combined via a 5-way DIN socket.

Wow and flutter measurement
Measurements without a test cassette first require recording the test frequency on to a tape inserted in the apparatus under test. The tracking tester supplies a stable reference frequency of 3000 Hz or 3150 Hz for this purpose. This frequency is obtained from a quartz crystal oscillator and a fairly complex filter that together guarantee a stable sine-wave signal with a distortion smaller than 1%.

After recording the test signal, the tape is rewound and played back. The wow and flutter measurement can then be carried out either with the reference tape or the previously recorded tape. Playing back the home-made recording will, in general, result in readings that are about 40% higher than those obtained with a proprietary reference tape. This difference is caused mainly by the fact that tracking fluctuations are also taken into account during the recording.

Toggle switch TEST/MEASURE is set to the MEASURE position, and the range switch to '1%'. The CALIBRATE control has no effect at this stage.

The wow and flutter percentage is read from the 1%-scale on the large moving-coil meter. High-quality and top-of-the-range recorders with very low wow and flutter may require the tracking tester to be set to the lower ranges of 0.3% or 0.1% full-scale deflection.

Speed deviation and drift
These measurements invariably require a proprietary test tape to provide a sufficiently accurate reference tone of 3000 Hz or 3150 Hz.

- For absolute speed measurement, the TEST/MEASURE switch is set to TEST.
- Set the range switch is to position 'Drift 5%'.
- Turn the CALIBRATE control until the meter indicates 0% at the centre of the scale.

With the TEST/MEASURE switch set to MEASURE, the meter indicates the absolute tape speed deviation in a range of 5%.

The service documentation with the

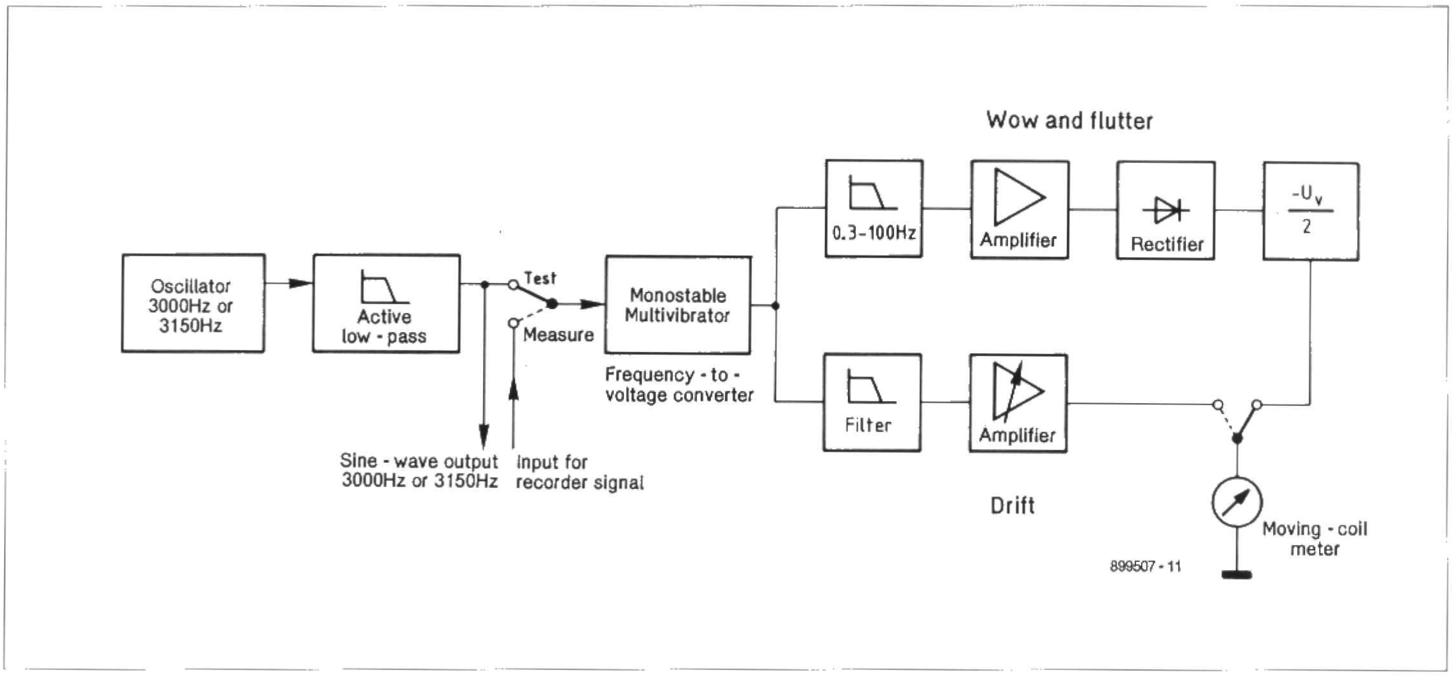


Fig. 1. Block diagram of the tracking tester.

tape recorder may be consulted to find out how the tape speed can be adjusted to obtain the lowest possible deviation (less than 0.5% is acceptable in most cases).

Following the absolute tape speed measurement, drift should be checked against the data supplied by the manufacturer. Rewind the test tape, and select the MEASURE mode. Calibrate the instrument for a reading of 0%. Play the test tape back to see how the long-term behaviour of the mechanical assembly affects average tape speed. Tape pull, determined by the varying effective diameter of the pulling reel is also a major cause of drift. Careful mechanical adjustments as outlined in the maintenance manual may give the required improvements.

Principle of operation

The principle of measurement used in the tracking tester is illustrated in the block diagram of Fig. 1. The quartz crystal oscillator and a number of dividers supply a rectangular signal at 3000 or 3150 Hz. A low-pass filter converts this signal into a sinusoidal shape with a distortion smaller than 1%. This is adequate for the measurements in question, because accuracy and stability are the prime aims.

When set to TEST, the mode selection switch supplies the generated reference signal to a monostable multivibrator which functions as a frequency-to-voltage (f-V) converter in conjunction with the filters that follow it.

The pulse ratio is such that the centre of the meter range can be set with the calibration control during drift measurements.

When the mode switch is set to the MEASURE position, the f-V converter is driven with the output signal of the recorder under test. When the absolute tape speed is higher than the standard speed, the input frequency of the converter is higher than the reference frequency (3000 Hz or 3150 Hz). This means that the MMV receives more trigger pulses within a unit of time, so that the low-pass filter at the output supplies a higher voltage. The needle of the moving-coil meter deflects to the right, and the relative deviation can be read from the scale as a percentage. Likewise, the needle deflects to the left of the centre when the tape speed is too low.

Wow and flutter measurements work on a similar basis. The only difference with drift measurement is the use of an additional rectifier that records short-term frequency changes which are the result of relatively fast tracking errors in the tape mechanism as discussed earlier. Since this measurement need not discriminate between positive and negative deviations, the full scale, starting at 0, is available for the read-out (a centre indication is not required). A differential amplifier is used for this purpose.

The circuit in detail

The circuit diagram is given in Fig. 2. The quartz crystal oscillator set up around gate N_1 supplies a 3.2768 MHz signal to divider IC_2, a Type CD4040. The divisor is either 1040 or 1092 as decoded by gates N_5, N_6 and N_7, and selected by switch S_1 (CCIR/DIN). In the first case, the output frequency at pin 14 of IC_2 is 3150 Hz, in the second case, 3000 Hz.

A first-order passive low-pass filter is formed by R_4, R_5, R_6 and C_3, and a third-order active low-pass filter by R_7, R_8, R_9, C_4, C_5, C_6 and opamp OP_1. The active filter is dimensioned for a roll-off frequency of about 3.5 kHz to achieve sufficient suppression of harmonics. A clean 3000 Hz or 3150 Hz sine-wave is available at pin 7 of OP_1. A further opamp, OP_2, raises this signal to a level of 0 dBm, or 775 mV_{rms} (2.2 V_{pp}) into 600 Ω. This level may be set by adjusting preset R_{11}.

The sinusoidal test signal is used for recording purposes by feeding it to the output terminals (pins 1 and 4 on the DIN socket, and the phono-type recording socket) as well as for use as an internal reference during drift measurements.

Toggle switch S_{2a} selects either the tape signal or the internal reference signal. Either of these is applied to the input of amplifier OP_3 via C_9 and R_{13}. Comparator OP_4 converts the output signal of OP_3 into a rectangular waveform.

The monostable multivibrator around N_3, N_4 and IC_6 is triggered at each rising pulse edge applied via C_{12} and R_{25}. Counter/oscillator IC_6 forms part of the MMV to ensure that the drift of the instrument itself can not affect the meter readings in the most sensitive range.

Each positive pulse transition at pin 12 of gate N_3 causes the output, pin 11, to change from high to low. This state is latched because of the bistable configuration of the two gates, and counter IC_6 is enabled via pin 12. The oscillator in the CD4060 starts to operate at a frequency determined by R_{28}, R_{29} and C_{16}. Ripple counter output pin 7 goes high after 8 clock cycles, which resets bistable N_3-N_4 via pin 2. The output of N_3 disables IC_6 and resets the MMV to its initial state.

Pin 3 of gate N_4 is high for the duration of the oscillator activity. This pulse is used for further processing, and is marked by high stability in respect of the pulse width.

That the instrument is capable of measuring tape tracking errors down to 0.1% is mainly by virtue of the third-order active filter around R_{27}, R_{41}, R_{42}, C_{20}, C_{21}, C_{22} and OP_6, in combination with first-order low-pass section R_{43}-C_{23}.

Opamps OP_7 and OP_8 are configured to provide further amplification, while S_{3b} forms the range selector.

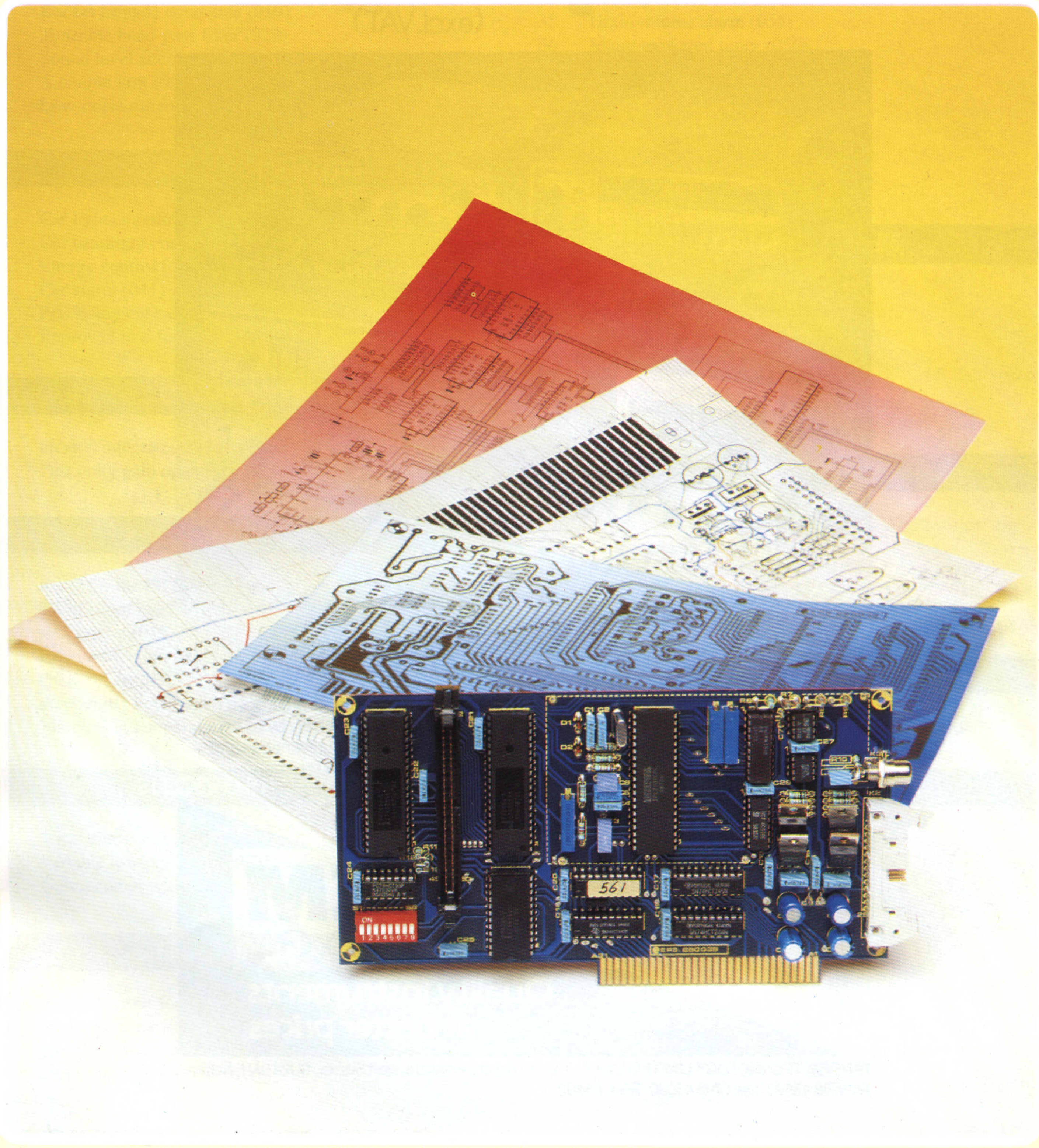
A peak rectifier is set up around OP_9 and D_{11}-C_{29}. The reference level of the rec-

THE ELECTRONICS MAGAZINE WITH THE PRACTICAL APPROACH

July/August 1989

Elektor Electronics

Supplement: a variety of construction projects from Audio & Hi-fi through Computers and Power supplies to Radio & TV.



CONTENTS

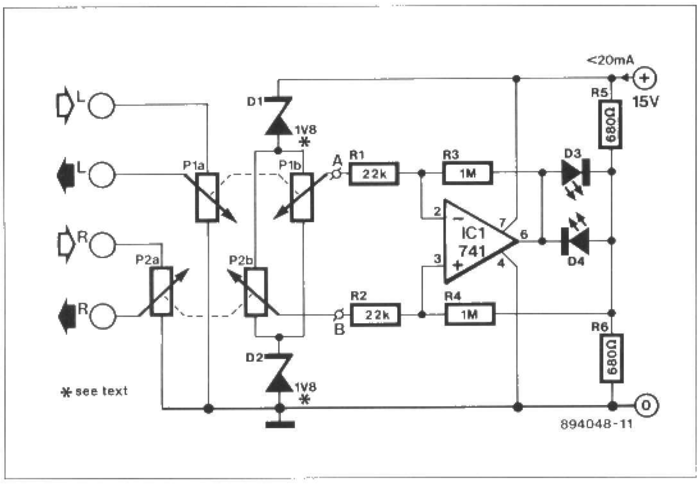
July/August 1989
Supplement

001 AUDIO & HI-FI

BALANCE INDICATOR

If your amplifier is fitted with two balance controls (as, for instance, the Elektor Electronics Preamplifier for Purists – Ref. 1), it actually offers you a balance control and a level control. A drawback of this is that it is quite difficult to set the balance properly. This may be obviated, however, by replacing the two mono potentiometers by stereo versions, P1 and P2 in the diagram.

One half of the pair, P1a and P2a, assumes the tasks of the removed components. The other half is connected in a bridge circuit. The voltage between the wipers of the potentiometers is then a measure of the balance between the two channels. The lower the potential, the better the balance. If you are interested in knowing the degree of unbalance, connect a centre-zero moving coil meter with a bias resistor between A and B. With this arrangement, zener diodes D1 and D2 may be omitted: they are necessary only with the LED indicator shown in the diagram to prevent the input voltage of the opamp getting too close to the level of the supply voltage.



The circuit around IC1 is a classical differential amplifier. Resistors R5 and R6 provide a virtual earth for the LEDs, which is necessary to ensure that in spite of the asymmetrical supply voltage a positive and a negative output is obtained.

Since the LEDs have been included in the feedback loop of the indicator, the circuit is pretty sensitive. At only 40 mV, that is, just one fourhundredth of the supply voltage, one of the LEDs begins to light already. The maximum current drawn by the LEDs is determined by the values of R5 and R6.

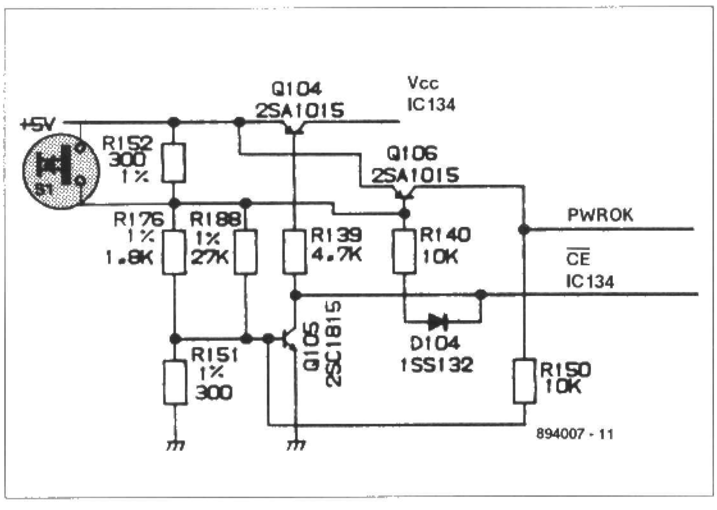
Ref. 1. "Elektor Electronics", October 1988.

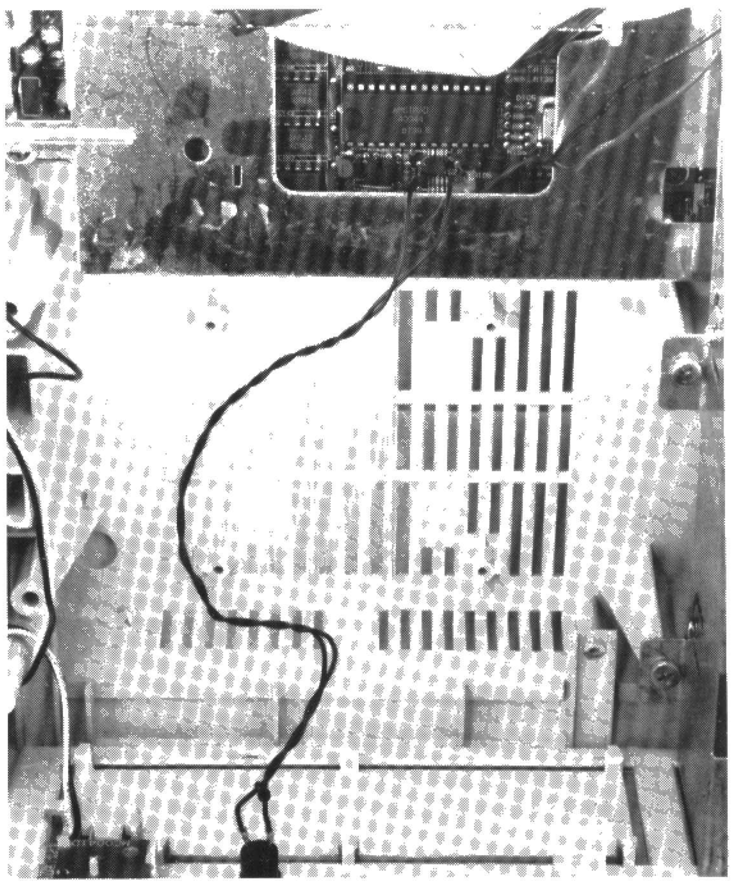
002 COMPUTERS

RESET FOR THE PC 1640

The PC1640 is one of Amstrad's popular and successful series of compact PCs. It has, unfortunately, a serious deficiency: there is no reset control. Luckily, it is not too difficult to fit this control retrospectively, since every PC, and the 1640 is no exception, has a reset circuit.

In the 1649, the circuit shown below monitors the level of the supply voltage. This voltage is sampled by potential divider R151–R152–R176–R188. If the supply is too low, the Q105, and consequently the Q104, switches off. Transistor Q106 is also provided with too little bias voltage to conduct. The enable input of the CMOS memory, IC134, is linked to the collector of Q105, so that when Q105 is turned off, the memory is disabled. At the same time, the supply to the memory is switched off since Q104 is off.





Only when the supply voltage is at the correct level does Q105 conduct. At the same time, Q104 and Q106 are provided with sufficient base current to switch on. This causes the supply to IC134 to be switched on, which results in the removal of the inhibit on the enable input. The processor receives a signal, PWROK, indicating that the supply level is all right, so that the reset cycle can start. The entire computer is initialized, while all I/O lines are set as required.

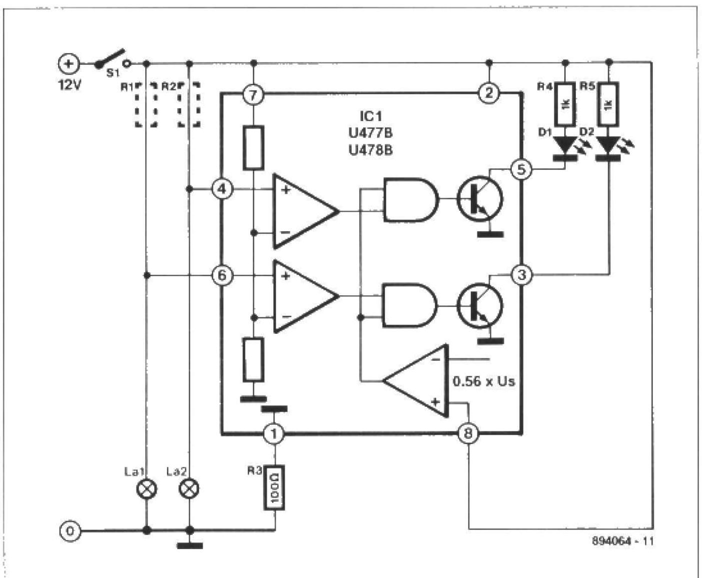
To fit a reset control, use is made of the PWROK signal in the following manner. If R152 is short-circuited, PWROK goes low and the processor is reset. Only when PWROK has gone high again will the processor start the reset routine. This is exactly the same cycle that occurs when the computer is restarted. In other words, all the reset control needs to do is to short-circuit R152.

The reset facilicity is incorporated fairly easily by fitting at some covenient place on the computer a simple push-button switch with a make contact. Connect the two terminals of the switch to the two ends of R152 (clearly marked on the PCB) via two short lengths of flexible wire and that's all.

003 CAR ELECTRONICS

CAR LIGHTS MONITOR

A defect car light is at best a nuisance and at worst a danger. Fortunately, most new cars are equipped with suitable monitors thast indicate on the dashboard whether a light is not working. There are, of course, millions of older cars that have no such sophistication and it is for these that the present monitor is intended.



Two special ICs are available from Telefunken that are designed for measuring the current through a light bulb. In practice, detecting whether a current flows through a bulb or no is a most suitable way of determining whether the bulb still works.

If a small resistance is connected in series with the bulb, a small voltage drop will develop across it when the bulb lights (R1 and R2 in the diagram). Each IC can cope with only two bulbs, so that per car three or four ICs are needed. The junction of the bulb and resistor is connected to one of the inputs (pin 4 or pin 6) of the IC. The potential across the resistor is compared in the IC with an internal reference voltage. Depending on which of the two ICs is used, the voltage drop must be about 16 mV (U477B) or 100 mV (U478B). This voltage drop is so small that it will not affect the brightness of the relevant bulb.

The value of the series resistor is determined quite easily. If, for instance, it is in series with the brake light (normally 21 W), the current through the bulb, assuming that the vehicle has a 12 V battery, is 21/12=1.75 A. The resistance must then be 16/1.75= 9 mΩ (U477B) or 100/1.75= 57 mΩ (U478B).

These resistors may be made from a length of resistance wire ((available from most electrical retailers). Failing that, standard circuit wire of 0.7 mm dia. may be used. This has a specific resistance of about 100 mΩ per metre. However, in most cars the existing wiring willl have sufficient resistance to serve as series resistor.

LEDs may be connected to the outputs of the IC (pins 3 and 5). These will only light if the relevant car light fails to work properly.

004 ELECTROPHONICS

VOCAL ELIMINATOR

Otherwise properly mixed sounds often suffer from a predominant solo voice (which may, of course, be the intention). If such a voice needs to be suppressed, the present circuit will do the job admirably.

The circuit is based on the fact that solo voices are invariably struated "at the centre" of the stereo recordings that are to be mixed. This means that the voice levels in the left- and right-hand channels are about equal. Arithmetically, therefore, left minus right is zero, that is, a mono signal without voice.

There is, however, a problem: the sound levels of bass instruments, more particularly the double basses, are also just about the same in the two channels. This is becausae on the one hand low-frequency sounds are virtually non-directional and on the other hand, the recording engineers purposely use these frequencies to give a balance between the two channels.

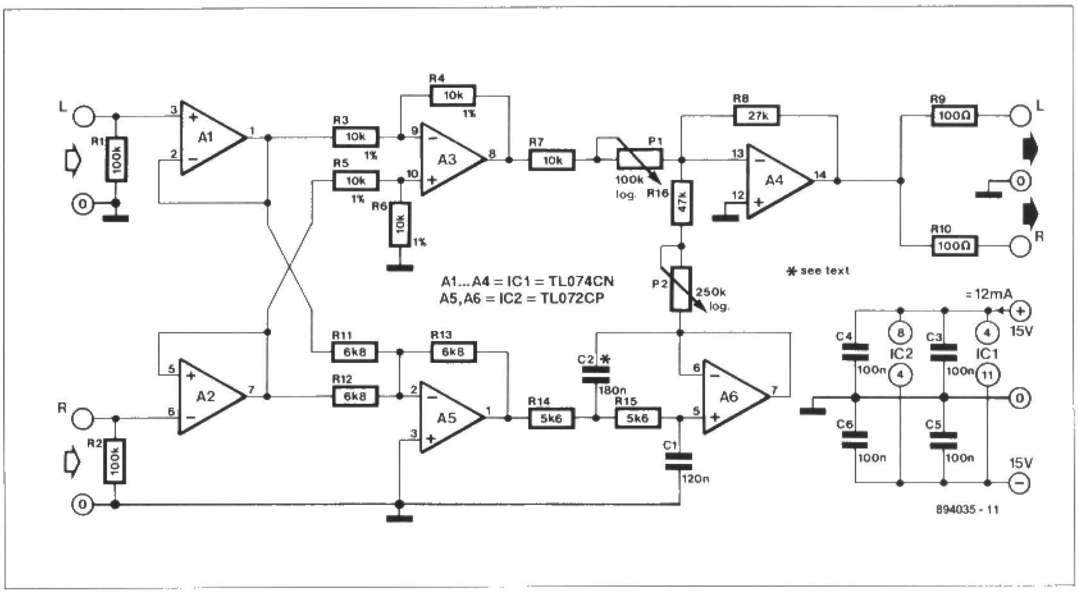
However, the bass instruments may be recovered by adding those appearing in the left+right signal to the left–right signal. The whole procedure is easily followed in the circuit diagram. The incoming stereo signal is buffered by A1 and A2. The buffered signal is then fed to differential ampolifier A3 and subsequently to summing amplifier A5. The latter is followed by a low-pass filter formed by A6. You may choose between a first-order and a second-order filter by respectively omitting or fitting C2. Listen to what sounds better.

The low-frequency signal and the difference signal are applied to summing amplifier A4. The balance between the two is set by P1 and P2 to individual taste.

You may have noticed that the circuit does not contain input

or output capacitors. If you wish, output capacitors may be added without detriment. However, the fitting of input capacitors is not advisable, because the consequent phase shift would adversely affect the operation of the circuit.

(A. Roelen)



005 GENERAL INTEREST

POWER BOOSTER FOR 7406/7407

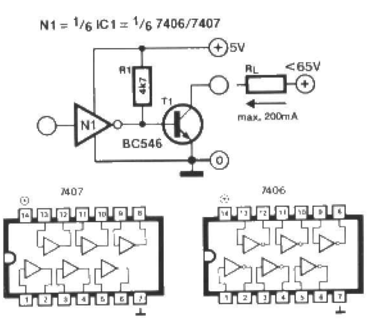
It often happens that a digital signal is required for controlling a relay, a stepper motor, or other kind of relatively heavy load. This makes it necessary for both the output current and the output voltage of the relevant device to be increased. Some logic devices are provided with constant-voltage open-collector outputs, but these are invariably restricted to 15 V or 30 V.

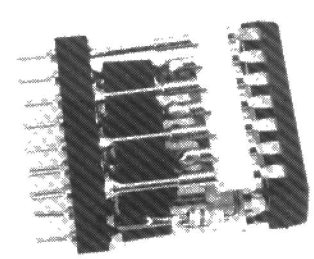
With a little dexterity, it is possible to provide a 7406 or 7407 with a dedicated open-collector output – see Fig. 1. If you aim for the construction in Fig. 2, the result will take not much more space than a standard 7406. Since the output stage inverts, it is necessary to use the non-inverting 7407 to obtain an inverted output.

The transistor must be chosen in accordance with the wanted output. For most general purposes, the BC546 is perfectly satisfactory (200 mA at 65 V).

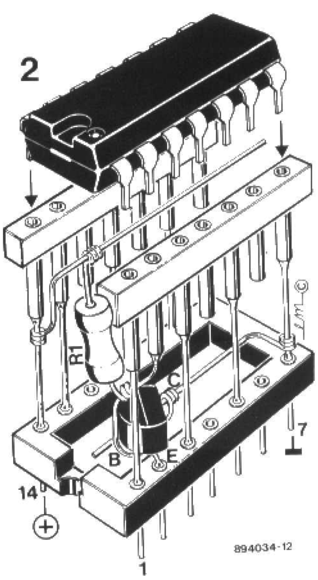
(A. Schaffert)

1





2



006 POWER SUPPLIES

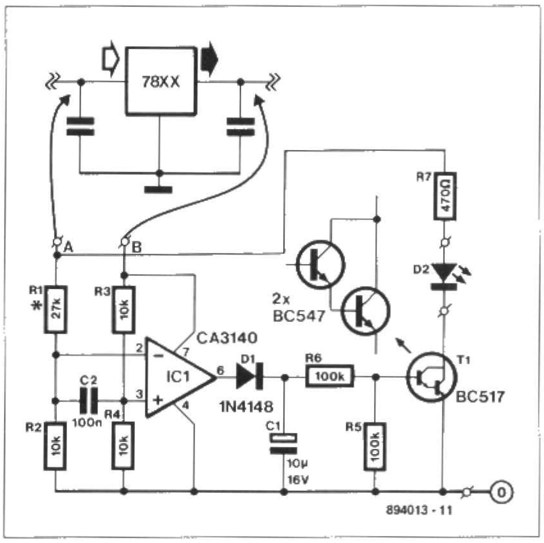
78XX MONITOR

When a voltage regulator is supplied from a mains adapter, it sometimes happens that its output is too low (because the output from the adapter is too low, or because the voltage has dropped to a low value owing to an overload). It is useful if a warning of that situation is indicated.

The proper operation of the 78xx series of regulators depends on the difference between input and voltage voltage, which must be not less than 3 V (worst case; many regulators are much better).

The voltage drop across the regulator is monitored by IC1. The input and output voltage of the regulator are supplied to IC1 via potential dividers. If the input voltage to the regulator is too low, IC1 goes high, which causes C1 to charge and this turns on T1, so that D2 lights. You may, of course, use a buzzer instead of D2 and R7. The charge on C1 ensures that the LED lights for at least 10 ms. This means that the circuit will react to even very short voltage drops at the input of the regulator. A large ripple that results in a too low input voltage is therefore clearly indicated. The circuit is based on the 7805; the value of R1 must be redimensioned for other members of the 78xx series by the following:

R1 = [(2dU / Ur) + 1] / R2



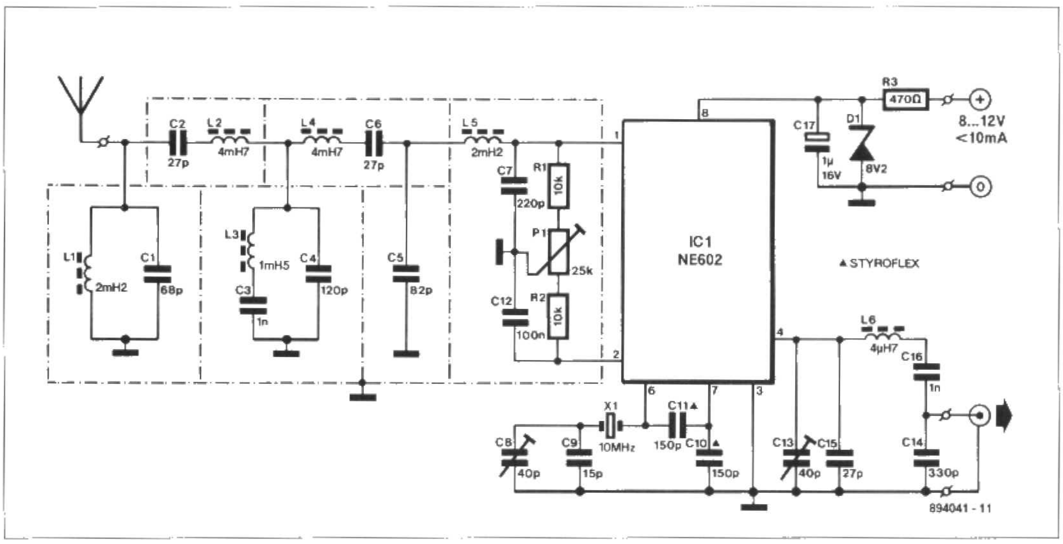
where dU is the voltage drop across the regulator and Ur is the characteristic output voltage of the regulator. It is necessary that dU is chosen somewhat larger than the actual minimum voltage drop across the regulator to prevent non-operation of the circuit. This is so because the minimum voltage drop across the regulator is constant when the device ceases to function properly until the input voltage returns to normal. In other words, the monitor must be able to react to a voltage drop that is slightly higher than the minimum voltage drop.

007 RADIO & TV

RADIO BEACON CONVERTER

The radio beacon band extends from 280 kHz to 516 kHz. Each beacon has its own characteristic AM modulated morse-coded call sign that is transmitted on a specific frequency. To be able to receive distant beacons, the aerial signal is passed through a band-pass filter that effectively suppresses long-wave and medium-wave signals. The filter also converts the aerial impedance, Zin, from about 10 kΩ to the inpout impedance of mixer IC1, which is about 1 kΩ.

The mixer adds or subtracts the received signal to/from the local oscillator signal, so that the beacon signal can be received on a normal short-wave receiver. The resulting frequencies lie in the range 9.72–9.484 MHz or 10.280–10.516 MHz.



In the construction of the converter some components must be surrounded by a metal shield as indicated by dashed lines on the PCB layout.

The circuit is aligned with the aid of an SSB receiver to which the output of the converter is connected. Tune the receiver to 10 MHz and adjust the oscillator frequency of the converter by means of C8 for zero beat. Next, detune the receiver slightly until a pleasant whistle is heard, which is adjusted for minimum level with the aid of P1. Finally, tune to a beacon transmitting at or about 300 kHz and adjust C13 for maximum sound output.

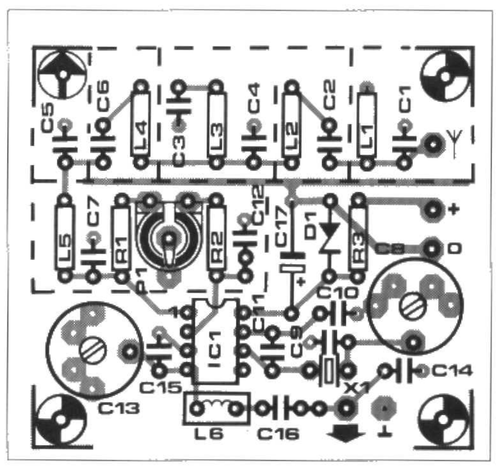
Parts list

Resistors:
R1; R2 = 10 kΩ
R3 = 470 Ω
P1 = 25 kΩ preset potentiometer

Capacitors:
C1 = 68 pF
C2; C6; C15 = 27 pF
C3; C16 = 1 nF
C4 = 120 pF
C5 = 82 pF
C7 = 220 pF
C8; C13 = 40 pF trimmer
C9 = 15 pF
C10; C11 = 150 pF styroflex
C12 = 100 nF
C14 = 330 pF
C17 = 1 µF; 16 V

Inductors:
L1; L5 = 2.2 mH
L2; L4; L6 = 4.7 mH
L3 = 1.5 mH

Semiconductors:
D1 = zener diode 8.2 V
IC1 = NE602



Miscellaneous:
X1 = 10 MHz crystal; 30 pF parallel resonance

SOUND LEVEL METER

Although the NE604 is, strictly speaking, intended primarily for h.f. applications, it may be used for a number of other purposes. One of these is a sound level meter for audio applications as presented here.

Use is made of the IC's signal-strength indicator that is based on an internal logarithmic converter. This enables us to obtain a linear decibel scale, so that the moving-coil meter shown in the diagram may be replaced by a digital instrument.

The signal source is assumed to be an electret microphone that converts ambient noise into an electriocal signal. Since this type of microphone normally contains a buffer stage, R7, R8 and C13 have been included to provide the supply voltage for this stage.

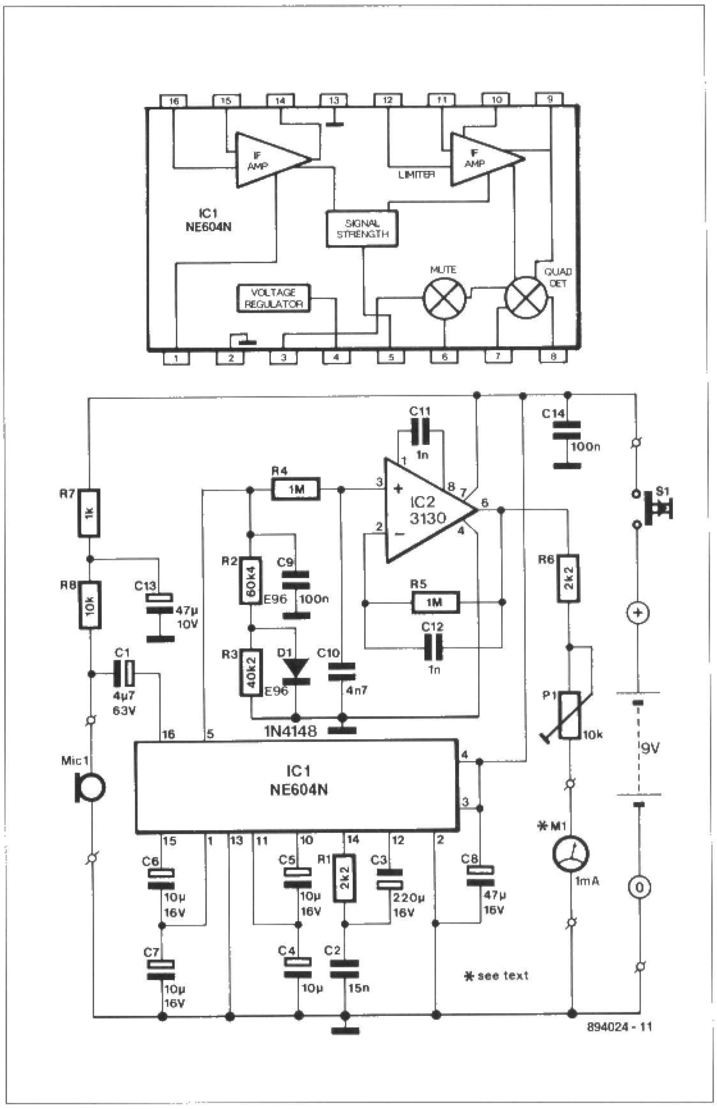
The NE604 delivers an output current (at pin 5) of 0–50 μA, which causes a potential difference across R2+R3 of 0–5 V. The range over which the relation between input and oujtput signal is logarithmic is, however, slightly smaller: about 0–4 V. This is equivalent to a sound range of 70 dB.

To compensate the effects of temperature changes, the required resistance of 100 kΩ is formed by two resistors, R2 and R3, and a diode, D1.

Any ripple remaining on the output voltage is removed by R4-C9-C10 before the output is buffered by IC2.

The indicating instrument, here a moving-coil meter, is connected to the output (pin 6) of IC2 via a series resistance, R6+P1. The preset is adjusted to give full-scale deflection (f.s.d.) for an output voltage of 4 V.

Calibrating the meter is a little tricky, unless you have access to an already calibrated instrument. Otherwise, if you know the efficiency of your loudspeaker, that is, how many decibels for 1 W at 1 metre, you can use that as reference. The scale of the meter can then be marked with the (approximate) value. In any case, the meter deflection must at all times be seen as an indication, not as an absolute value: it was not thought worthwhile to add a filter to the circuit to enable absolute measurements to be made.



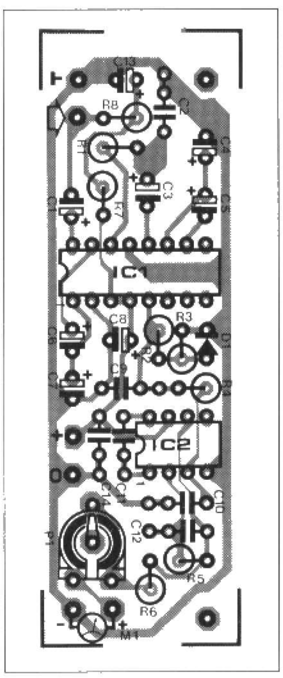
Parts list

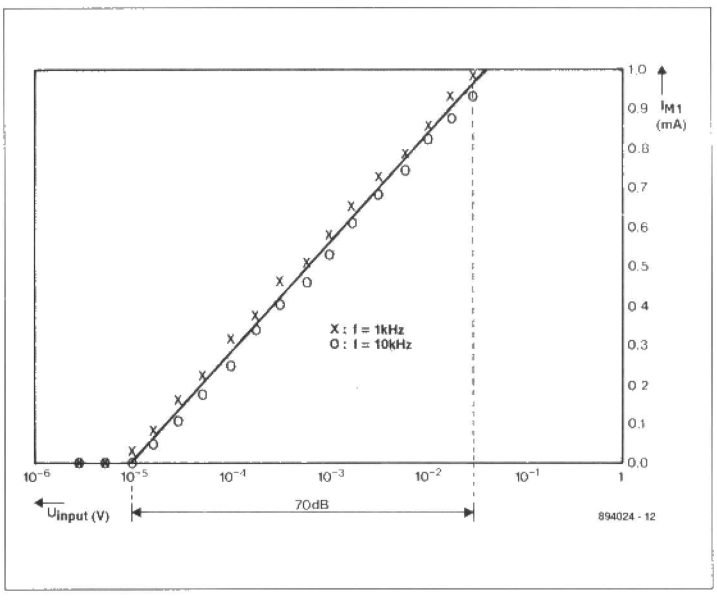
Resistors:
R1; R6 = 2k2
R2 = 60k4 (E96) or two 120 kΩ resistors in parallel
R3 = 40k2 or 39 kΩ in series with 1 kΩ
R4; R5 = 1 M
R7 = 1 k
R8 = 10 k
P1 = 10 k preset potentiometer

Capacitors:
C1 = 4μ7/63 V radial
C2 = 15 n
C3 = 220 μ/16 V radial
C4–C7 incl. = 10 μ/16 V radial
C8 = 47 μ/16 V
C9; C14 = 100 n
C10 = 4n7
C11; C12 = 1 n
C13 = 47 μ/10 V

Semiconductors:
D1 = 1N4148
IC1 = NE604
IC2 = 3130

Miscellaneous:
S1 = push-button switch with one make contact
M1 = moving-coil meter; 1 mA





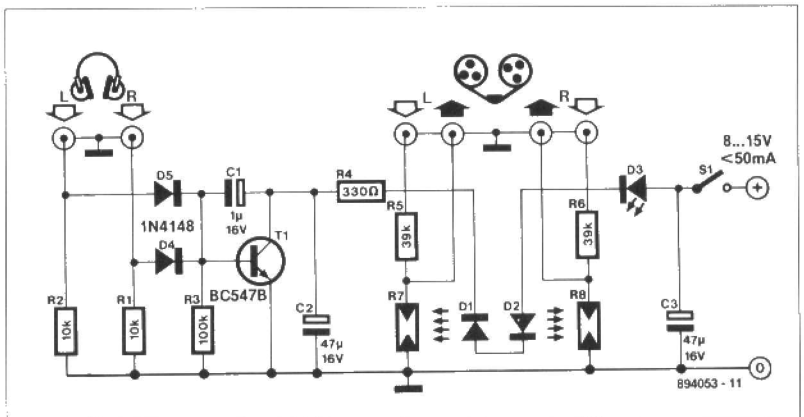
009 AUDIO & HI-FI

RECORDING CONTROL

The circuit presented here is intended as a recording level control for cassette recorders. It enables the reading on the VU meters to be kept out of the red sector without the necessity of the recorder's level control to be adjusted. This is particularly useful when speech is recorded.

The recording signal is controlled via two voltage dividers, each containing a light-dependent resistor – LDR. Unfortunately, these devices introduce slight distortion, but that is still better than a lot of distortion through overloading of the tape.

The circuit monitors the signal level at the headphone output of the recorder. The signal there is rectified (half wave) and then applied to T1. The speed with which T1 can react is determined by C1. As shown, the attack time is set at 3 ms. The release time of the circuit is determined by R4-C4 and is much longer than the attack time: about 100 ms.



The voltage across C2 determines to what extent the signal to be recorded is attenuated. The charging current through R4 depends on the voltage across the capacitor, so that the light intensity of the LEDs is also dependent on the signal strength.

The light from D1 and D2 falls on to R7 and R8 , which form part of the potential dividers that are added to the input circuit of the recorder.

The circuit is easily disabled by opening S1.

Owners of cassette recorder with separate recording and playback heads should bear in mind that the circuit does not work well if the internal monitor is switched on. In that situation, the signal takes a finite time to reach the headphone socket: too late for correcting the input signal level.

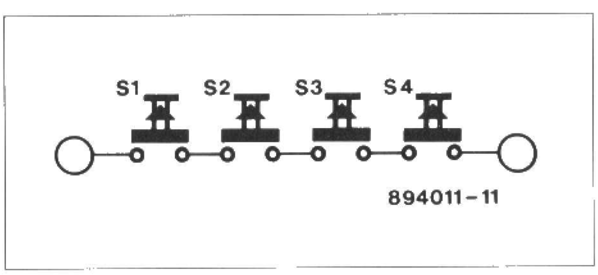
(S.G. Dimitriou and F.P. Maggana)

010 COMPUTERS

CHILD-PROOF RESET SWITCH

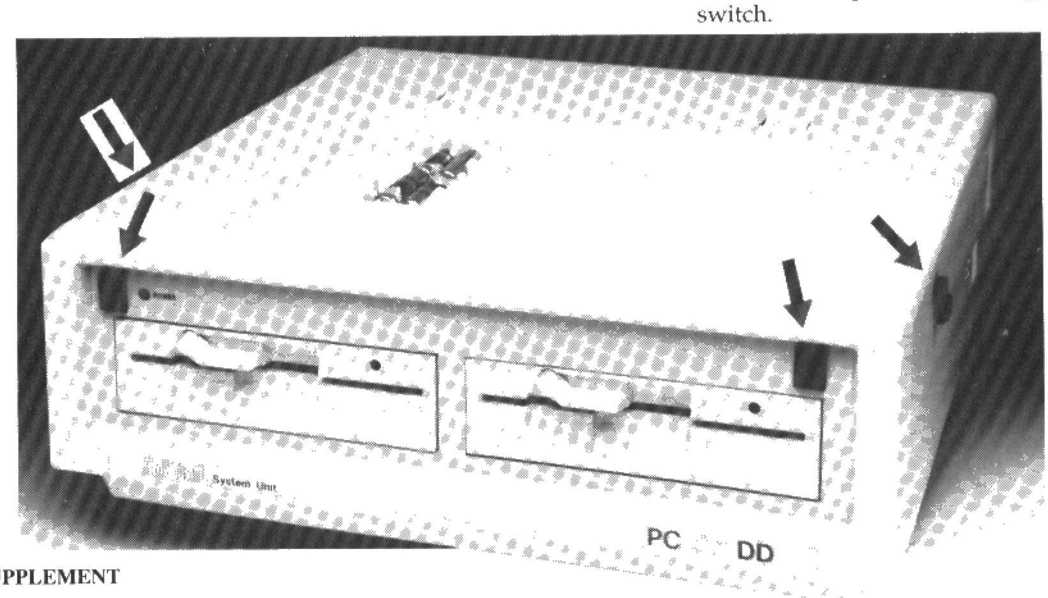
The reset switch on a computer is a very important control. If an operating instruction threatens to wreck the internal management of a computer, the reset button is often the only way of avoiding a possible disaster. On the other hand, it may also be the cause of a disaster. After all, one touch on it and hours of work may be negated in an instant, that is, if you do not save your work every fifteen minutes or so. None the less, anyone can have an accident, but it is particularly important that children or pets can not inadvertently operate the control.

The circuit proposed here should put an end to your worries in this respect. Instead of one reset switch, it is necessary to press four switches simultaneously. The chances of this happening by accident or child or pet are so small as to be negligible.



The four switches are placed in positions that make it impossible to operate them all with one hand. Instead, two of them can be operated with the fingers of one hand and the other two with the fingers of the other hand.

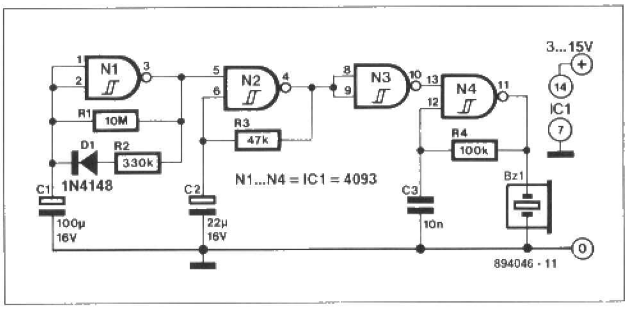
As shown, the four switches are connected in series and are intended to replace the existing switch.



011 GENERAL INTEREST

"ON" INDICATOR

Battery-operated equipment can work on one set of batteries for a long time nowadays. But if it is left on inadvertently, that 'long time' is over very quickly. Moreover, flat batteries are always found at the wrong moment. The circuit proposed here is a sort of aide-memoire. Every two minutes it emits 5–10 pips to indicate that the equipment is still switched on.



Basically, the circuit consists of three rectangular-wave generators and an inverter. The first of the generators is formed by N1 and provides a signal with a period of about two minutes and a pulse duration of around ten seconds. During those ten seconds, the second generator starts operating in a one-second rhythm. Thus, N2 outputs ten pulses every 2 minutes. That output is inverted so that N4, like N2, can only be enabled during the 10-second pulse train from N1. There is a difference, though: during those 10 seconds, N4 is enabled and inhibited ten times and this is what causes the pips.

Do not take the times and number of pulses too literally, because there are wide variances between ICs from different manufacturers. On the other hand, component values are not critical, so that it is fairly easy to adapt the circuit to personal taste or requirements.

The buzzer may be a standard Toko type or equivalent.

Finally, the current drawn by the circuit is negligibly small.

(R. Kambach)

012 POWER SUPPLIES

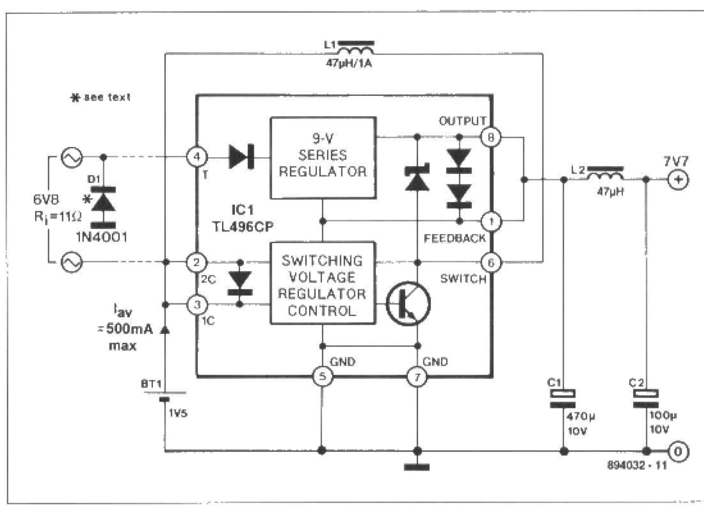
9-VOLT SUPPLY

The TL496 is an IC that is intended to produce a nominal 9-volt, that is 7–10 V, supply voltage from a variety of sources. This supply is perfectly adequate for many circuits.

The voltage source may be a 1.5 V battery, or two of them, or the secondary of a mains transformer.

To provide 7–10 V supply from these, the IC contains a series regulator and a switching regulator. The series regulator is connected actively like a transformer. A diode for half-wave rectification is also on board. When the series regulator delivers voltage at a satisfactory level, the switching regulator is inhibited. When, however, the output from the series regulator falls below requirement, the switching regulator comes into action. It is capable of generating a sufficiently high voltage level at the output from only one or two 1.5 V batteries.

The diagram shows how the TL496 must be connected for operation from one 1.5 V battery. For operation from two 1.5 V batteries, some modifications are necessary: pins 1 and 3 of the IC must be disconnected and left open, and the negative terminal of C1 must be disconnected from earth and instead be connected to pin 2 of the IC.



The circuit is eminently suitable for use with equipment that can also be supplied from the mains: normally, such equipment uses NiCd batteries. With the present circuit connected to the relevant terminals of the equipment, these batteries are recharged via D1. This occurs during the half cycle that is not used by the TL496. The charging current is determined by the internal resistance of the transformer (about 11 Ω, but note that this can not be measured with an ohmmeter). If you are in any doubt, connect a 10 Ω resistor in series with D1.

Finally, do not use D1 with dry batteries, because that would not do the batteries any good.

013 TEST & MEASUREMENT

ABS / RMS / LOG CONVERTER

The converter is specially intended for use in audio applications for which the IC used, a Type SSM2210 from PMI, is particularly suitable. The IC derives three voltages from the input signal: one corresponding to the absolute—ABS—value; one correspon-

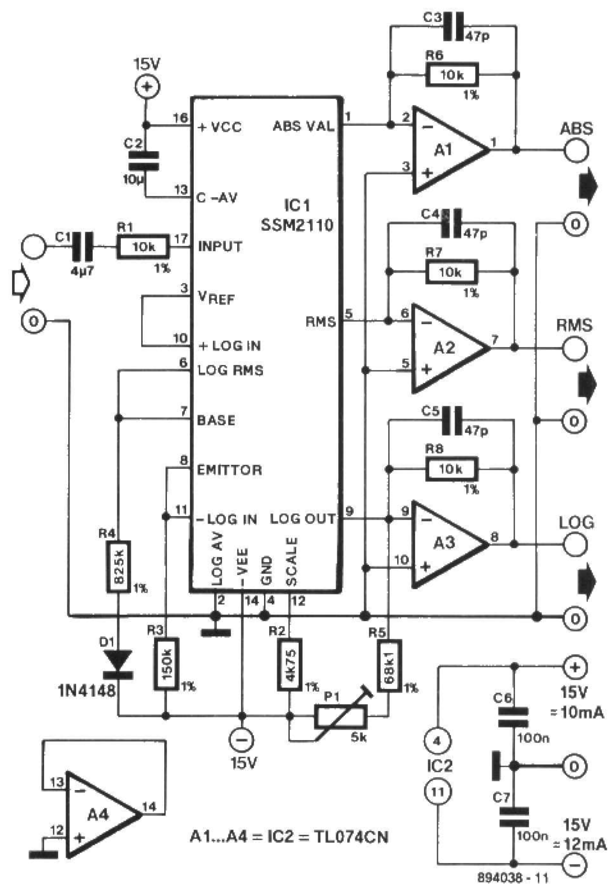
ing to the root-mean square—RMS—value; and the third corresponding to the logarithmic—LOG—value of the input current, since the IC operates with currents. This makes possible a wide dynamic range: something like 100 dB. Translated into terms of input current, that gives 3 mA (p-p) to 30 nA (p-p). This makes it vital that C1 is a type with a very small leakage current.

The input voltage is converted into current by R1. With a value as shown for this component, the nominal input level is 0 dBV with a reserve up to +20 dBV.

Like the input signal, the output signals are currents that are converted into voltages by opamps A1, A2 and A3, which are connected as current/voltage amplifiers.

To enable the decibel scale to be calibrated accurately, a preset potentiometer, P1 has been included in the A3 stage.

With component values as shown in the diagram, the output voltage from A3 varies at about 33 mV/dB. It should be noted, however, that the dynamic range of the LOG output at 80 dB is rather narrower than the maximum possible 100 dB.



014 AUDIO & HI-FI

4-CHANNEL MIXER

The proposed mixer is designed around four current-driven transconductance amplifiers contained in a Type SSM2024 from Precision Monolithics Inc—PMI. To obtain a low off-set and high control rejection, the four inputs should have an impedance to earth of about 200 Ω. These impedances are obtained from resistors R5–R8 that also form part of a potential divider at each input.

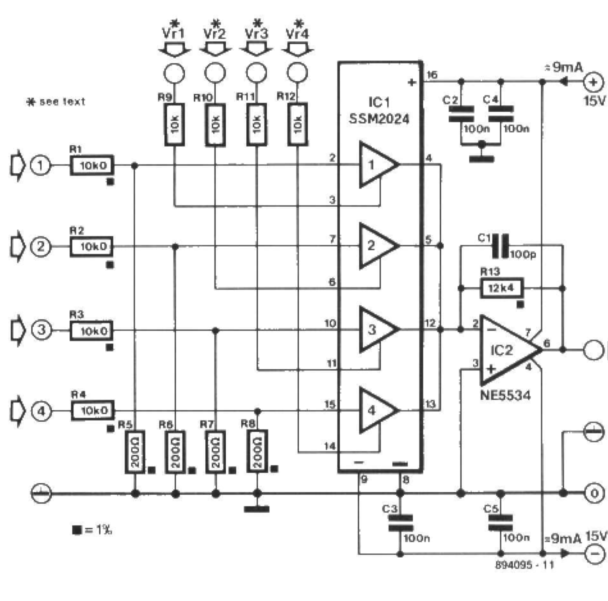
With values as shown in the diagram, the nominal input signal is 1 V (0 dBV). Distortion at that level is about 1% and at lower levels not more than 0.3%.

The amplification of the current-driven amplifiers—CCAs—is determined by the current fed into the control inputs. These inputs form a virtual earth so that calculating the values of the bias resistors (to transform the inputs into voltage-driven inputs) is fairly simple.

With a value for R1–R4 of 10 kΩ, the CCAs are switched off if the potential at the control inputs is lower than 200 mV. Maximum amplification is obtained at a drive current of 500 µA. The voltage at the control inputs is then slightly higher at 0.5 V, so that a maximum control voltage of 5.5 V is needed.

The output currents of the amplifiers are summed by simply linking the output pins (it is that simple with current outputs and completely in agreement with Kirchhoff's rules).

The current-to-voltage converter, IC2, translates the combined output currents into an output voltage. The value of R13 ensurest that the amplification of IC2 is unity.



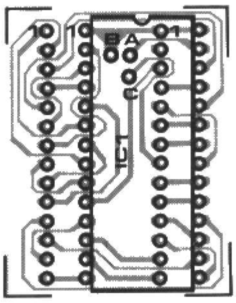
The current drawn by the mixer depends on the setting of the four CCAs and lies between 5 and 9 mA.

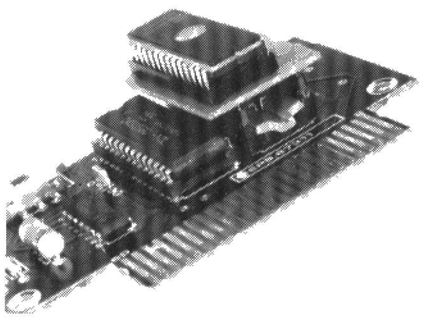
The signal-to-noise ratio of the mixer is about 90 dB, while the bandwidth is of the order of 130 kHz. The bandwidth is limited mainly by C1, which is essential to ensure good stability of the current-to-voltage converter.

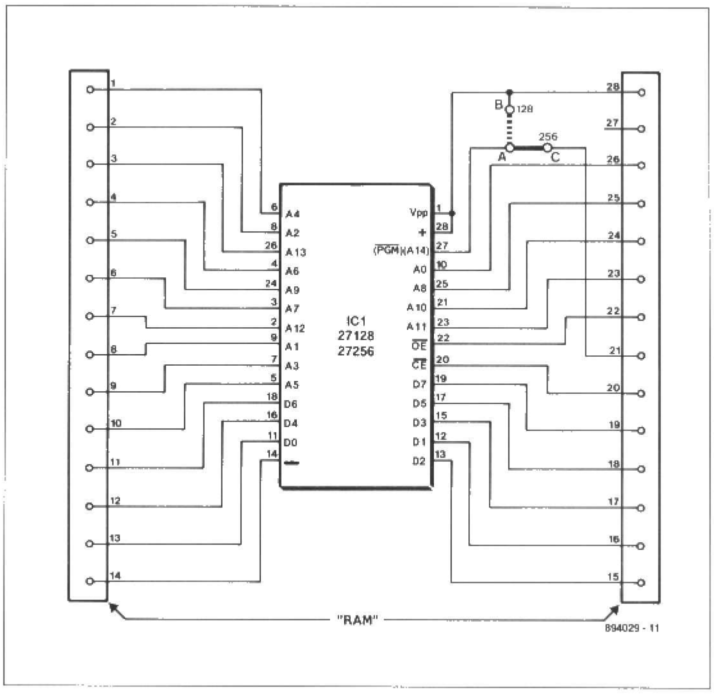
The SSM2024 operates satisfactorily with a supply voltage between ±9 V and ±18 V, but best results are obtained when it is ±15 V.

015 COMPUTERS

MSX EPROM







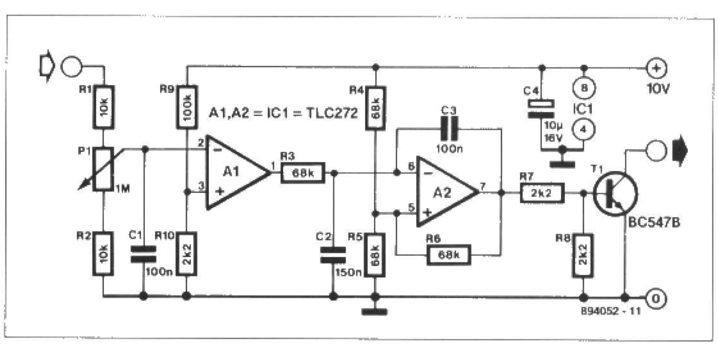
The "64 Kbyte RAM extension for MSX computers"(Ref. 1) can house EPROMs, but there is a small problem. To keep the board layout as simple as possible, the data and address lines on the extension board were not connected sequentially. With RAMs that does not matter too much and, in principle, not with EPROMs either, but their programming becomes a bit of a tangle. To prevent that, it is possible to use the auxiliary board shown here, which ensures that the address and data lines are connected to the correct pins of the IC. The board may be adapted to two types of EPROM by means of a wire link. With the link between A and B, a Type 27128 EPROM (16 K) may be used; if a 32 K EPROM is to be used, the link should be laid between A and C.

The two RAMs on the extension board each take two pages (a total of 32 K) in the address memory (pages 0/1 and 2/3 respectively). That means that when a Type 27128 EPROM is used, the data occurs twice on both pages.

Ref. 1. "Elektor Electronics", July 1988.

016 RADIO & TV

UNIVERSAL SQUELCH



This squelch circuit is simple, universally usable and provides a large enough gain to enable it being incorporated in the automatic-gain control (AGC) circuit of a variety of receivers.

The input signal derived from the AGC circuit in a receiver is attenuated by network R1-R2-P1. The signal at the wiper of P1 is taken to the inverting input of opamp A1, which is connected as a comparator. The non-inverting input is provided with a reference voltage of 200 mV by potential divider R9-R10. The output signal of A1 is applied to a Schmitt trigger circuit, A2, via low-pass section R2-C2. This filter ensures that small noise and other interference signals do not affect the correct fucntioning of the squelch.

Capacitor C3 removes the steep skirt of the output signal of A2, which renders the operation of the AGC rather more pleasant to the ear. The output of A2 is then taken to the base of output transistor T1 via potential divider R7-R8.

The open-collector output of the squelch may be used to suppress the audio frequency output of the receiver.

Since the squelch draws only a small current, less than 10 mA, its incorporation into an existing receiver should not present any problems as far as the power supply is concerned.

(R. Lalic)

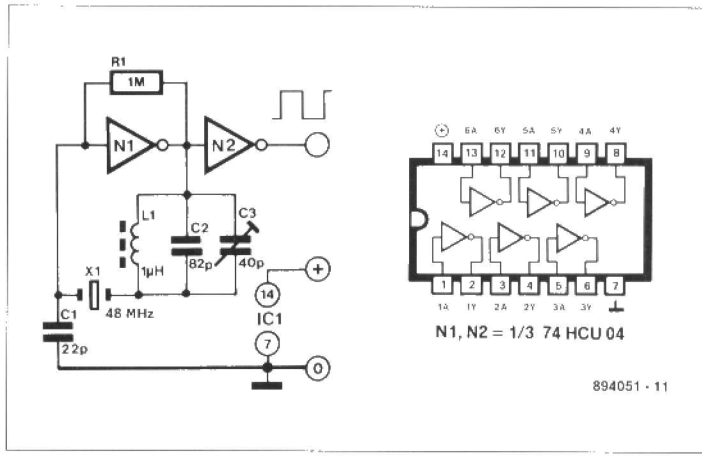
017 TEST & MEASUREMENT

48 MHZ CMOS OSCILLATOR

Crystal oscillators using digital gates generally do not generate frequencies above 30 MHz, because the necessary crystals are not allowed to oscillate on their fundamental frequency.

In the oscillator shown here, the crystal is forced to oscillate on the third harmonic since it is connected in series with a parallel tuned circuit that resonates at the fundamental of 16 MHz. For digital applications, the output of N1 may be enhanced by the use of a second inverter.

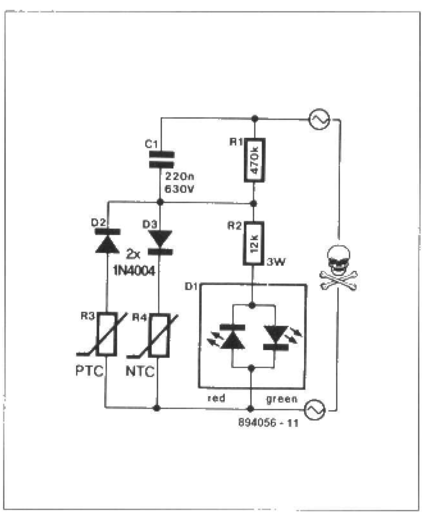
The circuit will operate satisfactorily only if non-buffered CMOS devices are used. Gates from the HCU family will enable operation up to 60 MHz.



018 GENERAL INTEREST

SIMPLE TEMPERATURE INDICATOR

For the absolute measurement of temperatures a thermometer is indispensable. There are, however, many situations where an absolute value is not needed and a relative indication suffices.The getting hot of an electric drill or vacuum cleaner may be indicated to the user by the lighting, or changing of colour, of a simple indicator light. It would be a further advantage if on such equipment a green light would indicate that all is well as far as temperature is concerned. As the temperature rises, the light should change colour slowly to indicate that the equipment is getting too hot.



The proposed circuit does this and has the additional advantage that it does not need a separate low-voltage supply: it works direct from the mains. The indicator proper is a two-colour LED, D1, while the sensor is a combination of a negative temperature coefficient (NTC) and a positive temperature coefficient (PTC) resistor, R4 and R3 respectively.

At a relatively low temperature, the value of R3 is low and that of R4 is high. During the positive half cycle of the mains voltage, a voltage will exist across R3-D3 that is sufficiently high to cause the green section of D1 to light. The value of R3 has been chosen to ensure that during the negative half cycle of the mains voltage the potential across it is too low to cause the red section of D1 to ligh.

If he temperature rises, the value of R4 diminishes and that of R3 rises. Slowly but surely the green section will light with lesser and lesser brightness, while at the same time the red section lights with greater and greater brightness until ultimately only the red section will light.

Resistor R2 and capacitor C3 ensure that the current drawn by the LEDs does not become too large. This arrangement keeps the dissipation relatively low.

Both R3 and R4 should be of reasonable dimensions, something like 6 mm in diameter – not less. At a temperature of 25 °C, the NTC must have a value of 22–25 kΩ and the PTC one of 25–33 Ω.

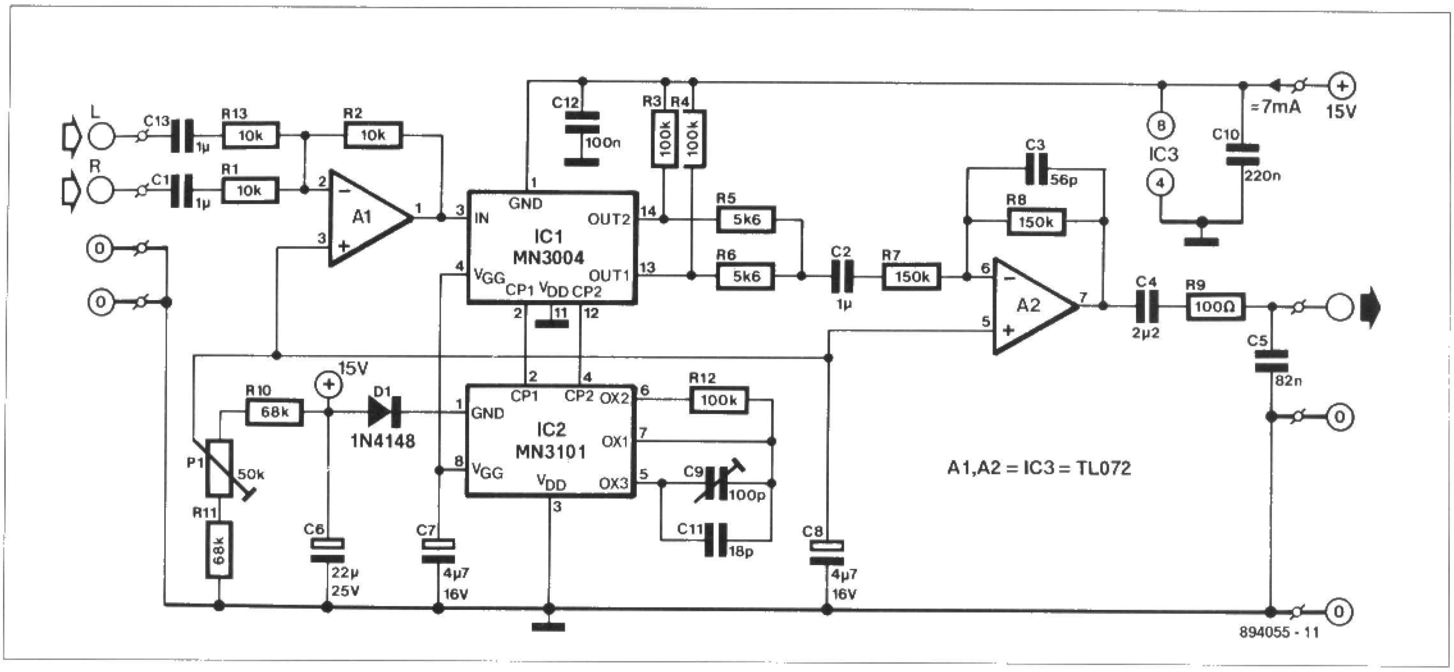
The circuit should be treated with great care since it carries the full mains voltage.

019 AUDIO & HI-FI

BUCKET BRIGADE DELAY LINE

Although bucket-brigade devices (BBDs) are not in the same limelight as many other components, they do exist and are used. One of them, Type MN3004, consists of 512 capacitors and switching transistors. The IC is perhaps best described as an analogue shift register. The capacitors are provided by the drain-gate capacitance of the transistors.

Samples of an analogue signal taken at the input appear 256 clock pulses later at the output. The clock pulses are obtained



from an associated IC, a type MN3101. This IC also provides a supply voltage, Ugg, for the source followers in IC1. Note that the connections of the supply voltages to IC1 and IC2 have not been drawn incorrectly: Vdd must be negative with respect to earth (GND pin).

The MN3101 also makes it possible to construct an RC oscillator. Trimmer C9 enables varying the clock frequency and thus the delay time. It is also possible to connect an external oscillator to pin 7 (but note that pins 5 and 6 should be left open).

The clock frequency of the delay line may lie between 10 kHz and 100 kHz. With component values as shown in the diagram, it is about 60 kHz.

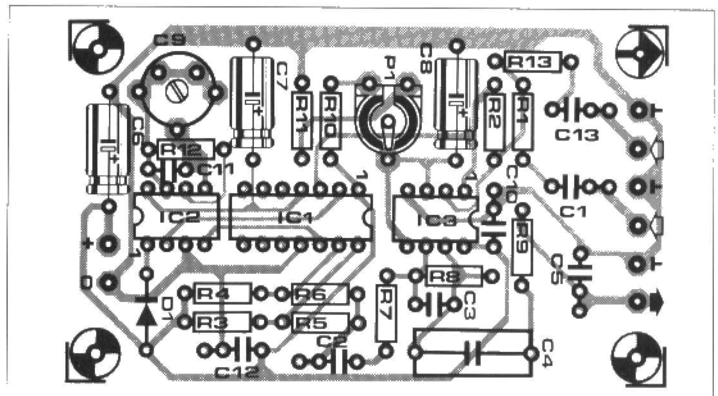
Maximum dissipation of the clock generator is about 200 mW, depending on the capacitive load of the memory. If more memories are used, it is possible to reduce the dissipation by connecting resistors in series with pins CP1 and CP2. Lowering the clock frequency will also reduce the dissipation.

The delay time of the circuit is equal to half the number of capacitors divided by the clock frequency: with values as shown, it amounts to 2.56–25.6 ms. The bandwidth of the delay line is roughly 0.33 times the clock frequency. Thus, with a clock frequency of 60 kHz, the bandwidth is 20 kHz and the delay more than 4 ms. In the prototype, these values yielded a signal-to-noise ratio of more than 70 dB and a distortion not exceeding 0.3% (at 1 kHz; 0 dBV). The current drawn was just over 6 mA, but this may increase to 14 mA with increased clock frequencies.

Apart from the delayed input signal, the output signal contains mixing products of the input and the clock. In the prototype (circuit as shown), with a clock of 60 kHz, these products were suppressed in the audio range by 60 dB (R8-C3 and R9-C5). It is, however, advisable when a lower clock is used to filter both the input and the output with a filter of at least the fourth order. Try to minimize the distortion by adjusting P1.

The delay line should find application in units for echo, tremolo, vibrator, chorus, reverb, and so on.

Another possibility is its use in a compressor to suppress, or at least attenuate, the short periods of overdrive. To that end, delay the input signal prior to the voltage-controlled amplifier (VCA) and drive the circuit that provides the control voltage for the VCA direct. In that application, the delay should be at least equal to the attack time of the compressor.



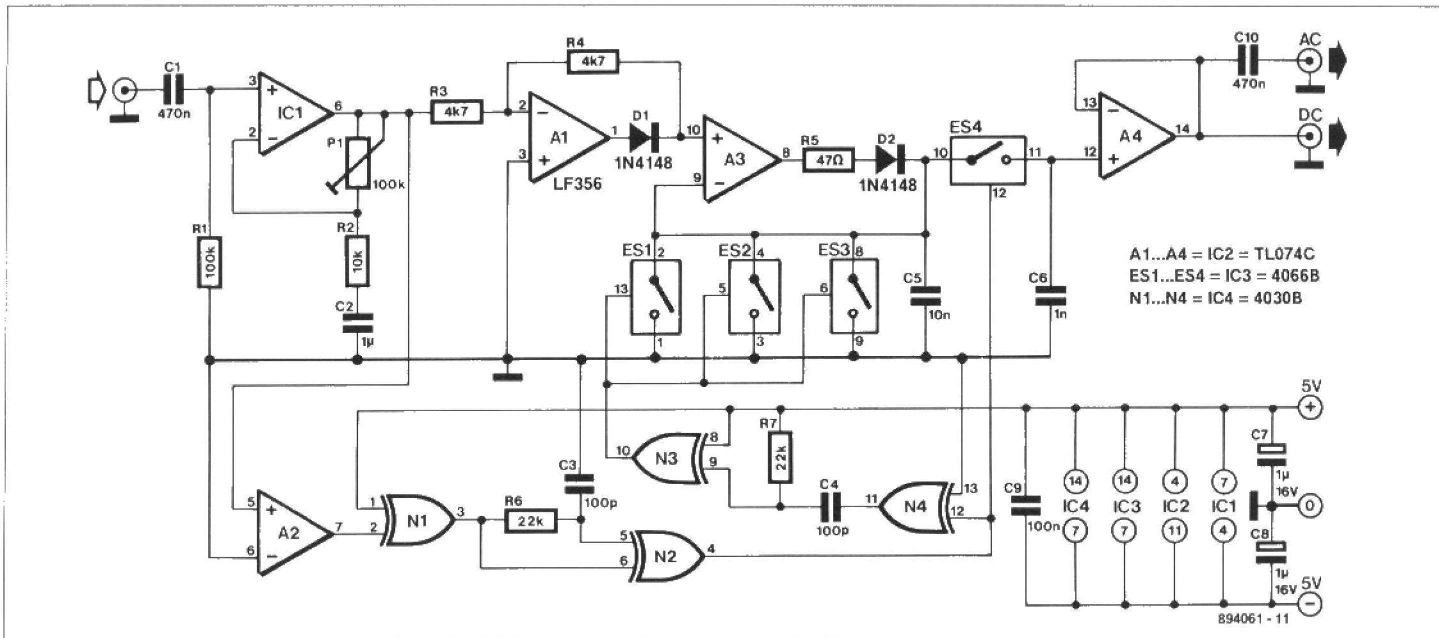
Parts list

Resistors:	C4 = 2µ2
R1; R2; R13 = 10 k	C5 = 82 n
	C6 = 22µF; 25 V
R3; R4; R12 = 100 k	C7; C8 = 4µ7; 16 V
R5; R6 = 5k6	C9 = 100 p trimmer
R7; R8 = 150 k	C10 = 220 n
R9 = 100 R	C11 = 100 p
R10; R11 = 68 k	C12 = 100 n
P1 = 50 k preset	
PC5; PC6 = PC2	Semiconductors:
	D1 = 1N4148
Capacitors:	IC1 = MN3004
C1; C2; C13 = 1 µF	IC2 = MN3101
C3 = 56 p	IC3 = TL072

020 RADIO & TV

FAST ENVELOPE SAMPLER

When a signal is being amplitude-modulated, constant monitoring is necessary to ensure that the maximum modulating frequency does not exceed half the carrier frequency. The fast envelope sampler presented here is basically an advanced AM



demodulator that can be used with modulations where the well-known diode detector with LF filter can not. Phase errors that occur with diode detection are absent in this circuit.

Amplifier IC1 is a buffer with variable (1–11) AC amplification. Opamps A1 and A3 form an active half-wave rectifier that charges capacitor C5 up to the maximum signal level in a half period.

During the zero crossing, detected by A2, the network around N2 generates a short pulse. This pulse ensures that the potential across C5 is applied to C6 via electronic switch ES4. After ES4 has been opened, so that the charge is stored safely in C6, capacitor C5 discharges via parallel-connected switches ES1–ES3 (which are actuated via the network around N3 and N4).

The output of the sampler is in accordance with the potential * ABOVE COLUMN RH.

*across C6. Basically, the circuit is frequency-independent since the clock signal is derived from the carrier. This is the reason that the sampler may be used in such diverse applications as satellite facsimile, radio receivers and speech processors.

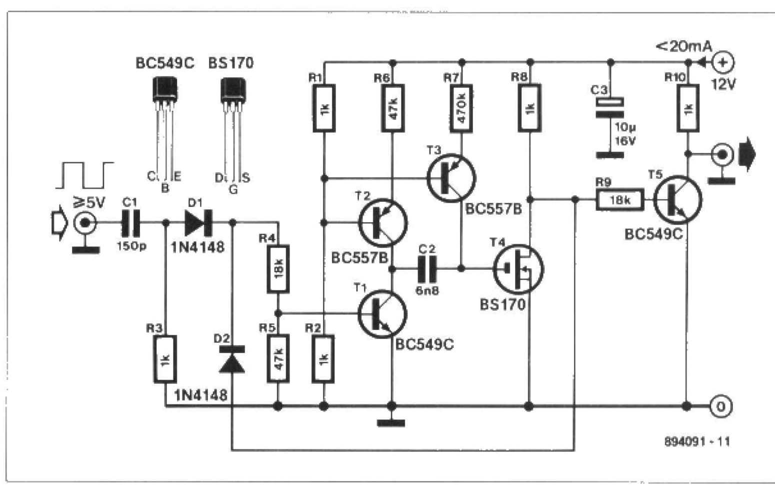
The circuit may also be used to good effect in an AGC loop, because the classical problems regarding attack and delay associated with diode detection and LF filter systems do not occur in this type of demodulator.

Since in some applications a direct voltage is needed, for instance, in an AGC loop, the proposed circuit has a DC as well as an AC terminal.

The sampler is intended for operation from a ±5 V supply and draws a current of about 20 mA.

021 TEST & MEASUREMENT

NOISE GENERATOR



The noise generator presented here provides constant noise energy over its bandwidth, resulting from the non-linear behaviour of its switching components, more particularly T4. It is very useful for measurements where limited noise bands are required. Varying the ratio R6:R7 and the clock frequency enables the generated noise to be adapted to specific requirements.

Transistors T2 and T3 are current sources. The current through T2 is about ten times the level of that through T3.

Assuming that T4 is on and that the clock input is low, T1 is off and C2 discharges. The capacitor is pulled to about half the supply voltage by the two current sources. When that state is reached, stability ensues because the potential then present at the gate of T4 keeps the FET switched on.

When the clock goes high, T1 is switched on so that C2 is connected between the gate of T4 and earth. Since C2 is only partly charged, the FET is switched off. Transistor T1 is kept switched on by OR gate D1-D2 so that the clock pulses are blocked. Capacitor C2 then charges via T3 until the potential across it becomes high enough to switch on T4. Transistor T1 is then switched off and the circuit is ready to receive another clock pulse (or rather a leading edge of one).

Since it is not known when the clock pulse arrives, it is not known to what potential C2 will be discharged by T2 (and countered by T3). It is therefore also not known how long it takes T3 to recharge C2. It follows that it is then not known when the next clock pulse arrives. In other words, the pulse width of the output signal is varying constantly, which is characteristic of a noise signal.

The frequencies contained in the noise signal are limited by the clock signal (higher frequencies than the clock can not occur, although there are harmonics) and the maximum charge and

022 AUDIO & HI-FI

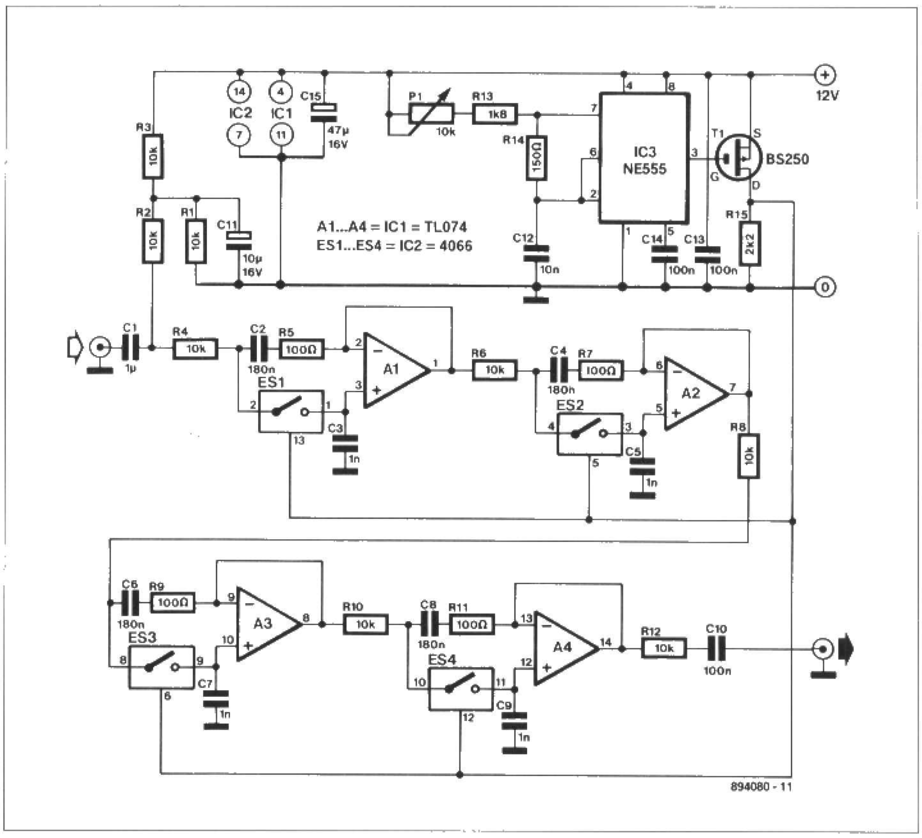
TUNEABLE BAND-PASS FILTER

One of the difficulties in the design of higher-order tuneable band-pass filters is achieving correct tracking of the variable resistors in the RC networks. The use of switched capacitor networks can obviate that difficulty as is shown in this filter.

The filter may be divided roughly into two stages: an oscillator that controls the electronic switches and the four phase-shift networks that provide the filtering proper.

The oscillator, based on a 555, generates a pulsating signal whose frequency is adjustable over a wide range: the duty factor varies from 1:10 to 100:1.

Electronic switches ES1–ES4 form the variable resistors whose value is dependent on the frequency of the digital signal. The operation of these switches is fairly simple. When they are closed, their resistance is about 60 Ω; when they are open, it is virtually infinitely high. If a switch is closed for, say, a quarter of the time, its average resistance is therefore 240 Ω. Varying the open:closed ratio of each switch varies the equivalent average resistance. The switching rate of the switches must be much greater than the highest audio frequency to prevent audible interference between the audio and clock signals.



The input signal causes a given direct voltage across C1, so that the opamps may be operated in a quasi-symmetric manner in spite of the single supply voltage.

The direct voltage is removed from the output signal by capacitor C10.

The fourth-order filter in the diagram may be used over the entire audio range and has an amplification of about 40, although this depends to some extent on the clock frequency. The bandwidth depends mainly on the set frequency.

The circuit draws a current of not more than 15 mA.

023 CAR ELECTRONICS

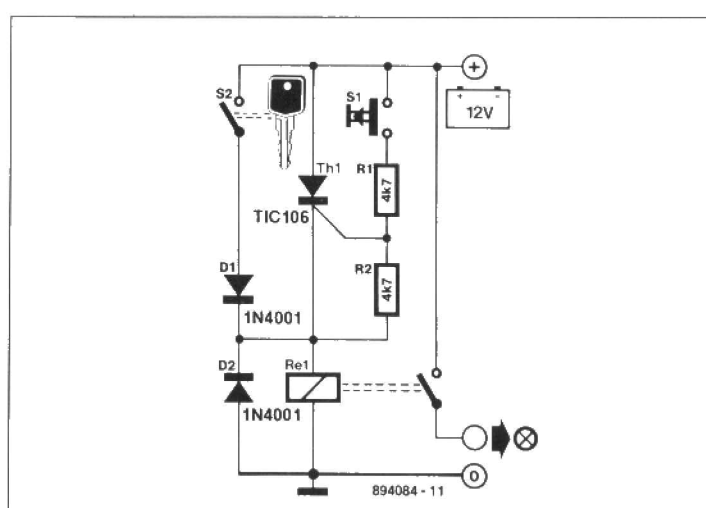
CAR HEADLIGHT CONTROL

It is an annoying fact that you normally only realize that you have left your car headlights on when you want to restart the car only to find that the battery is flat. One of the possible ways of preventing this happening is offered by the present control.

The circuit does not provide a warning but an action: when you switch off the ignition, relay Re1 is de-energized and the headlights are switched off, unless you deliberately decide otherwise. That decision is made possible by switch S1, which, when operated, triggers silicon-controlled rectifier Th1 so that Re1 is energized. Note that this is possible only when the ignition switch, S2, is off, otherwise the voltage across Th1 is so low, owing to shunt diode D1, that it can not be triggered. Since, however, the headlights should not normally be switched on when the ignition is off, in most cases S1 will be used only rarely and the switch may then well be omitted altogether.

Relay Re1 should be a standard 12 V car type with contacts that can switch up to 25 A.

(H. Huynen)



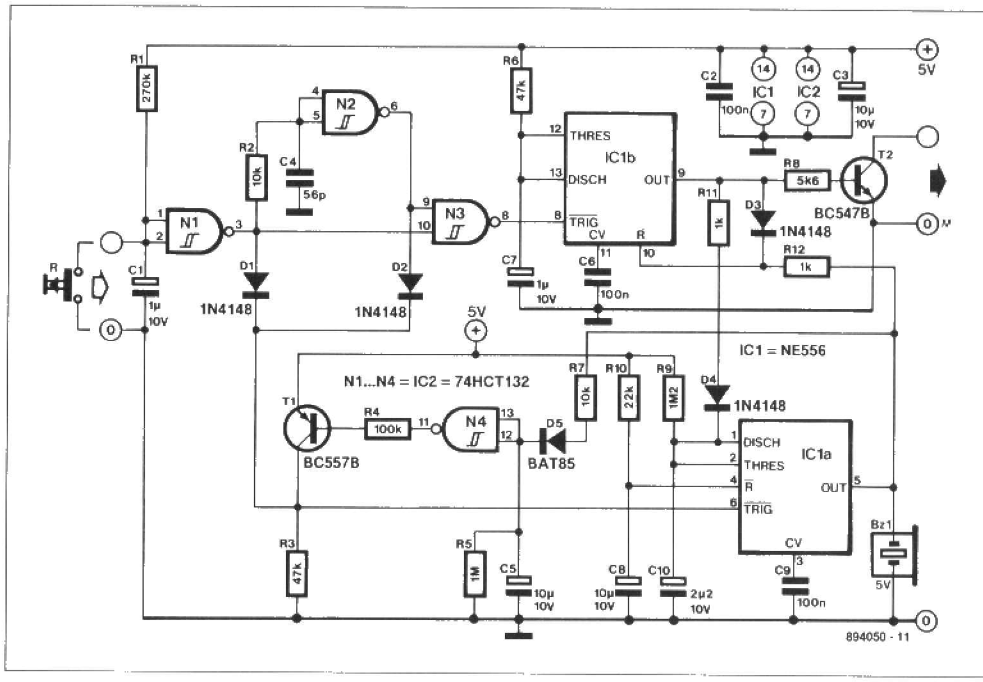
024 COMPUTERS

RESET PROTECTION

Most advanced computer programs include preventative measures against (possibly) hasty instructions. Responses such as "are you sure?" and "do you really want to quit?" are familiar to most of us. However, even the cleverest program can not prevent the inadvertent operation of the reset switch and the consequent result of lost data and improperly closed files that cause waisted clusters on the hard disk.

The location of the reset switch on the front panel of many computers asks, of course, for inadvertent operation. Clearly, some means of reset protection is no luxury.

Normally, the reset switch is connected to the mother board of the computer via two wires. One of these is at earth potential and the other is linked to the reset circuit. The protection, whose circuit is shown in the diagram, is inserted between the reset switch and the mother board. The earth connection of the computer must be linked to terminal M of the protection circuit. The protection circuit may draw its power from the computer supply.



When the circuit has been fitted, operation of the reset switch will not result in an immediate restart of the computer. Instead, a buzzer will sound to alert you to the reset operation. The buzzer is actuated for four seconds by monostable IC1a, which is triggered by the reset switch. During those four seconds, the output, pin 5, of IC1a ensures that the reset function, pin 10, of IC1b is disabled. When then the reset switch is operated again, monostable IC1b will be triggered and this will start the reset procedure. Transistor T2 is then switched on for half a second and the buzzer is deactuated via R11 and D4.

The circuit around T1 and N4 ensures that IC1a can accept trigger pulses again ten seconds after the mono time of IC1b has lapsed. This arrangement prevents, say, children operating the reset switch.

025 COMPONENTS

BICMOS INTEGRATED CIRCUITS

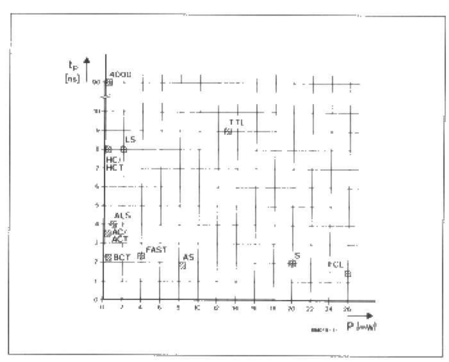
BiCMOS devices combine bipolar and CMOS technologies, providing the best features of both: the fast performance and the 48/64 mA output drive of bipolar devices and the low power consumption of CMOS devices.

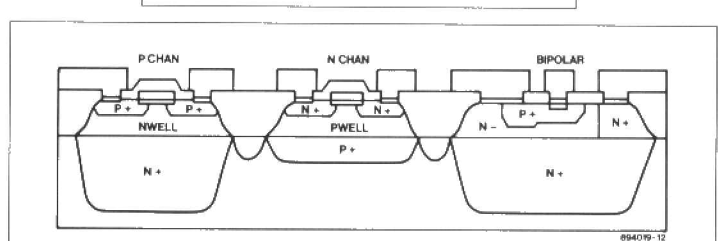
In active mode, BiCMOS devices operate at about half the supply current of their pure bipolar equivalents. When disabled, they may reduce power consumption by up to 90%. Since at present most BiCMOS ICs function as bus interfaces (that are normally disabled), the result may be a system IC-power saving of up to 25%.

Moreover, BiCMOS devices use the typical 0.3–3.5 V TTL voltage swings at their output rather than the larger GND-to-Vcc swings of CMOS devices. This smaller voltage swing reduces the overall effects of transient voltage noise produced during the simultaneous switching of multiple outputs.

A number of integrated circuits are available in BiCMOS technology, including transceivers with registers, pipeline registers, 8/9/10-bit registers, latches and parity bus transceivers. These give the designer additional means of reducing power consumption without compromising advanced performance.

(Source: Texas Instruments)

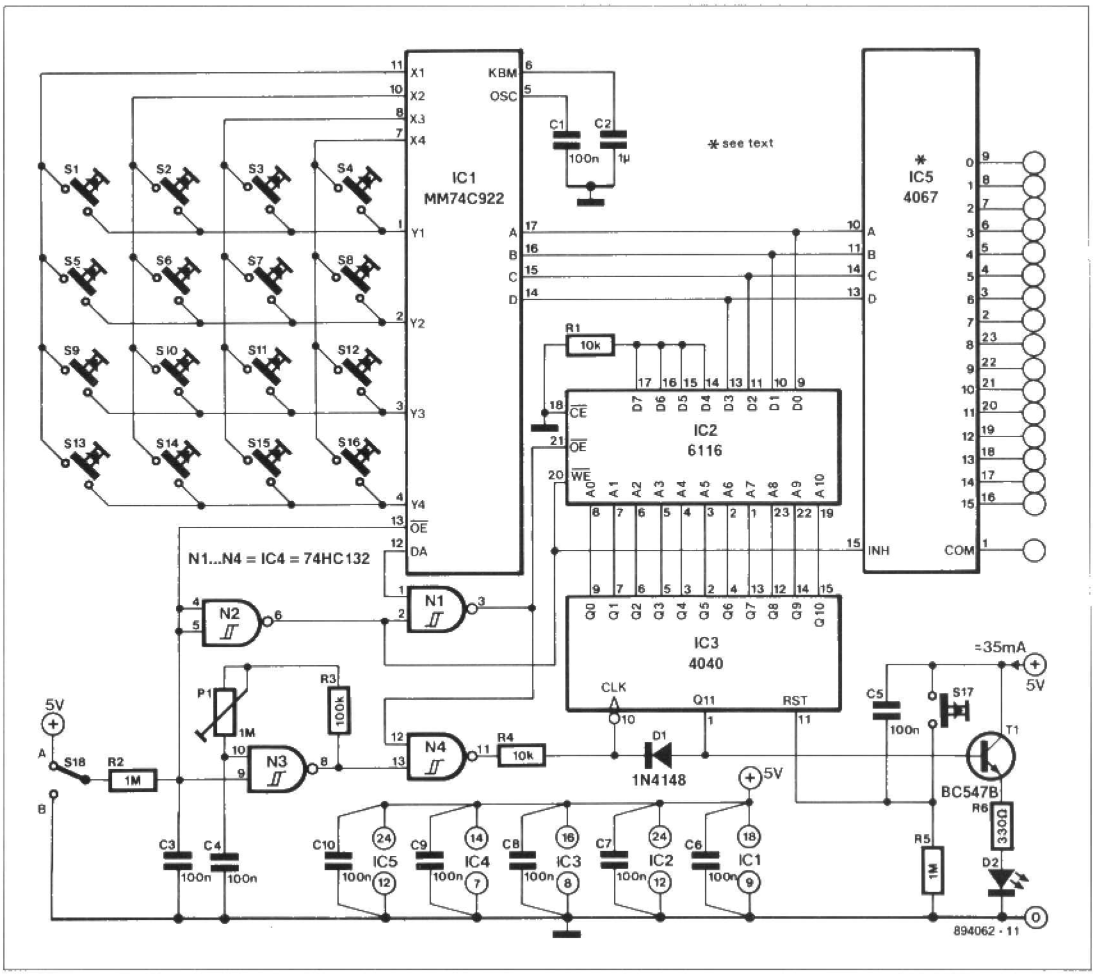




PROGRAMMABLE SWITCH

The programmable switch may be used, for instance, to simulate a data stream or, as shown, to control an analogue multiplexer, IC5. The multiplexer may be used as the basis of a programmable oscillator.

The circuit is based on National Semiconductor keyboard decoder Type MM74C922 (IC1). This device is intended to read a 4×4 matrix keyboard in a simple and fast manner. Apart from a 4-bit output, the IC also has a DATA AVAILABLE output, DA, which is high as long as one of the keys is pressed. The data associated with the last pressed key will remain on the output, even after the key has been released. The speed with which the keyboard is scanned is determined by C1. When this capacitor has a value of 100 nF, as shown in the diagram, the scanning rate is 600 Hz. The anti-bounce period of the keyboard is determined by C2: with a value as shown, the period is about 10 ms (rule of thumb: C2≈10C1).



The programming of the RAM, IC2, is fairly simple with the aid of IC1. First, the DA output of the decoder is inverted by N1 to enable it being used as the WRITE pulse. After the key has been released, the RAM is disabled (DA=1, $\overline{WE}$=0), and the address counter, IC3, is clocked on by one address. Since the counter is a Type 4040, it reacts to trailing edges and this means that WE has to be inverted again (by N4). It is not possible to use the original DA signal, since this could jeopardize the timing and cause non-defined states. The delay times of the gates ensure that all processes take place in correct sequence. Further delay is provided by the combination of R4 with the capacitance of the clock input.

The programmed data is read with the aid of a separate clock signal generated by an oscillator based on N3. The speed at which the data is read from the RAM is set by P1. Gate N3 is actuated by setting S18 in position A. Contact bounce here is prevented by the combination of R2 and C3, and the hysteresis of N3 and N2.

When S18 is in position A, gates N1 and N2 ensure that the RAM is in the read mode (WE=1). At the same time, N2 arranges that the data lines of the RAM are connected as outputs. Gate N4 ensures that either DA or the output of N3 is used as the clock for counter IC3. Furthermore, S18 also arranges the disabling of the data outputs of IC1 to prevent a bus conflict (the RAM now provides the data).

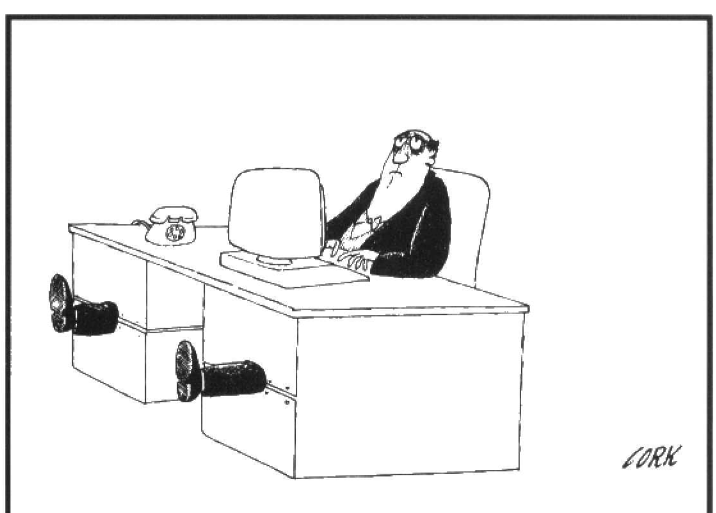
Pin 1 (Q11) of counter IC3 is connected to the clock input, pin 10, via OR network R4-D1 to ensure that the counter stops after one cycle. The stopping is indicated by the lighting of D1. Switch S17 enables the counter to be reset; it may also be used to set the counter to zero during programming.

When, during programming, the highest counter position is reached, the clock for the counter is disabled. Continued pressing of the keys will cause overwriting on to the last RAM address.

To obtain a defined counter state on switch-on, network R5-C5 provides a power-on reset of the counter.

Only one half of the RAM is used, since four bits suffice for the proposed circuit. It is, therefore, possible to use a RAM with nibble configuration or use two keyboard encoders.

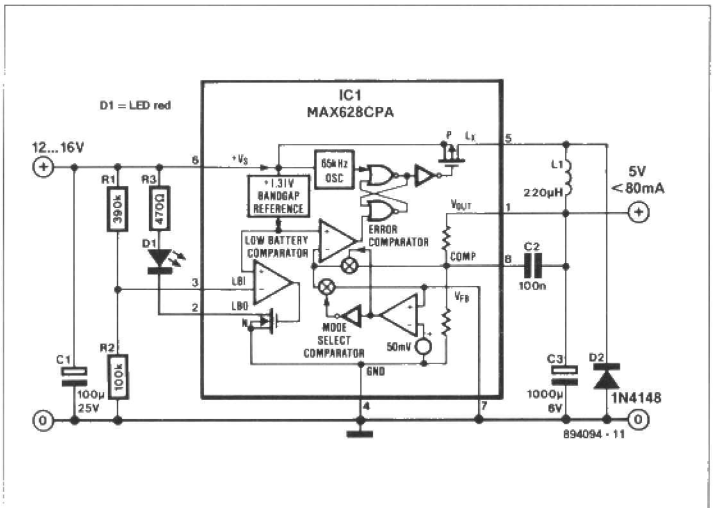
Finally, the use of 16 separate switches may make the circuit rather expensive: it is far cheaper to use a second-hand 4×4 matrix membrane keyboard.



027 POWER SUPPLIES

SWITCH-MODE VOLTAGE REGULATOR

Switch-mode power supplies offer the user the benefit of a much greater efficiency than obtainable with a traditional power supply. The switch-mode regulator presented here has an efficiency of around 85%.



An input voltage of 12–16 V DC is converted into a direct voltage of exactly 5 V. The use of a Type MAX638CPA enables the design and construction of the regulator to be kept fairly simple: only nine additional components are needed to complete the circuit.

Resistors R1 and R2 are used to indicate when the battery voltage becomes low: as soon as the voltage on pin 3 becomes lower than 1.3 V, D1 lights. With values as shown for the potential divider, this corresponds to the supply voltage getting lower than about 6.5 V.

The output of the IC is shunted by a simple LC filter formed by L1, C3 and D2.

The oscillator on board the IC generates a clock frequency of around 65 kHz and drives the output transistor via two NOR gates. The built-in error detector, the 'battery low' indicator or the voltage comparator can block the clock frequency, which causes the transistor to switch off.

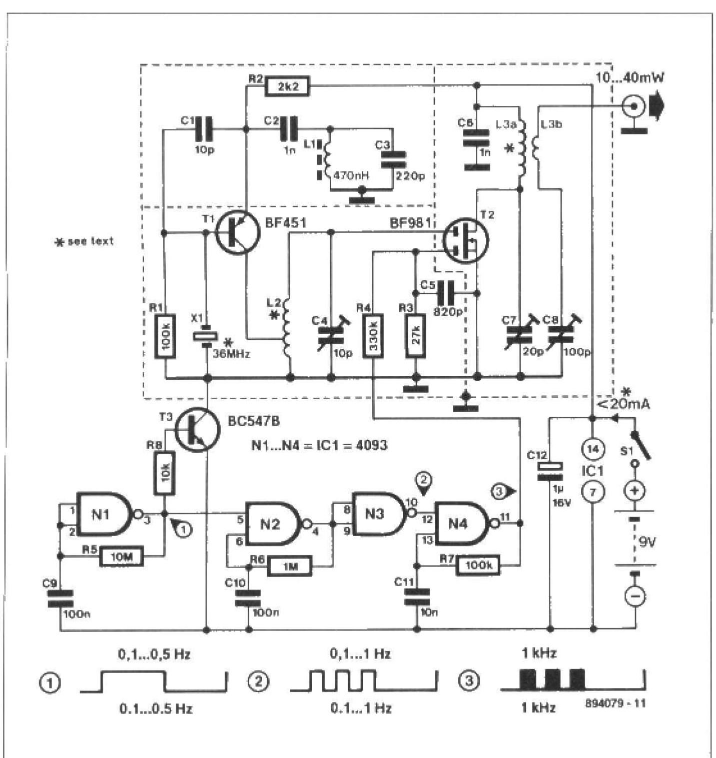
The IC compares the output voltage of 5 V with a built-in reference (FET). Depending on the load, the FET will be switched on for longer or shorter periods. The maximum current through the FET is 375 mA, corresponding to a maximum output current of 80 mA

028 RADIO & TV

2-METRE TRANSMITTER

The transmitter was designed primarily for use by radio amateurs as a radio beacon and as such it provides a good quality signal free of unwanted harmonics.

Transistor T1, in association with crystal X1, operates as a 36 MHz oscillator. Filter L1-C3 obviates any tendency of the circuit to oscillate at 12 MHz (the fundamental frequency of the xrystal).



Circuit L2-C4 is tuned to the fourth harmonic of the oscillator signal (=144 MHz). This signal is fed to the aerial via a buffer stage consisting of T2, a double-gated FET. The (amplitude) modulating signal is applied to the second gate of the buffer. The output power of the transmitter has been kept low, about 10–40 mW.

The modulating signal is generated by N1, an oscillator that switches the transmitter on and off via transistor T3. The switching rate lies between 0.1 Hz and 0.5 Hz.

When the output of N1 is low, T3 is switched off, and the transmitter is inoperative because the supply is disabled. When the output of N1 is high, T3 is on and the transmitter operates normally.

The digital pattern at the gate of T2 shapes the modulating signal. Gate N2 generates a square wave at a frequency of 0.1–1 Hz. As long as the output of T3 is high, N4 oscillates at a frequency of about 1 kHz. At the relevant gate of N2 there is, therefore, a periodic burst-signal at a frequency of 1 kHz, and this signal is used to modulate the transmitter.

The digital pattern at the relevant gate of T2 may be varied to individual requirements by altering the values of the feedback resistors in the digital chain.

The transmitter is calibrated by setting trimmers C4, C7 and C8 for maximum output power.

Inductors L2 and L3 are wound from 0.8 mm dia enamelled copper wire: L2 = 5 turns with a tap at 1 turn from ground; L3a = 3 turns and L3b = 2 turns. The coupling between L3a and L3b should be arranged for maximum output power.

The circuit draws a current of only 20 mA, enabling the transmitter to be operated from a 9-V PP3 battery for several hours.

029 TEST & MEASUREMENT

METER-SCALE MAGNIFIER

The resolution of moving-coil meters is generally no better than 1%, because the scale normally is given no more than 100 marker lines for the given dimensions: more lines might detract from the legibility. Most digital meters have a resolution of 0.05% or better. The resolution of moving-coil instruments may be improved in two ways: physically enlarging the scale, which is possible only in the factory, or electronically enlarging the scale, which is what this article is all about.

The circuit divides the scale into five sections, each of which is then extended over the full scale. This therefore gives a five-fold improvement in resolution.

The input signal (200 mV = full-scale deflection – f.s.d.) is amplified by IC1 to a value of 2.5 V. The amplified signal is fed to four comparators that divide the input signal into segments of 40 mV. Which segment is indicated on the meter-scale is shown by the LEDs.

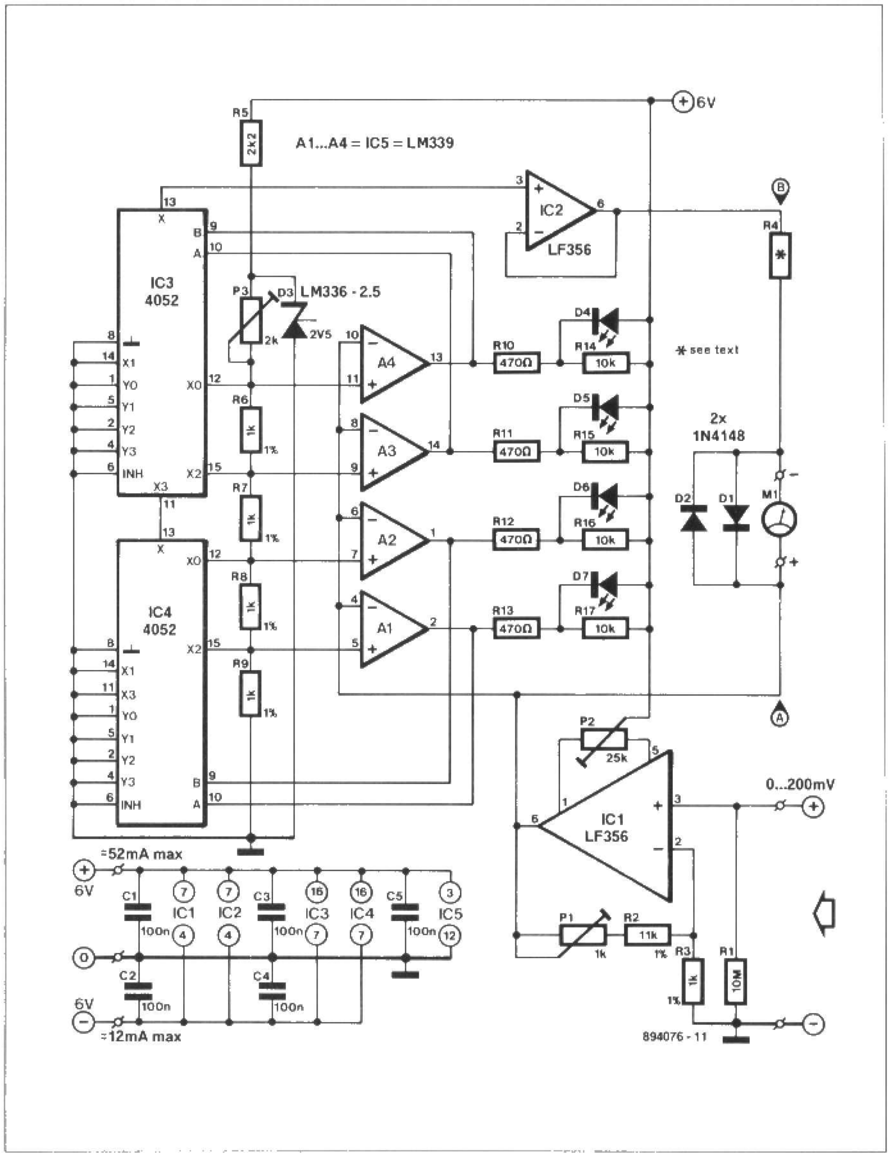
The LEDs are shunted by a 10 kΩ resistor, so that the open-collector outputs of the comparators have a well-defined pull-up resistance.

The outputs of the comparators drive two multipliers that, depending on the magnitude of the input signal, provide a direct voltage to buffer IC2. This direct voltage is always a multiple of 0.5 V (since the output of IC1 is 5 × 0.5 = 2.5 V).

There then exists a potential between A and B that is equal to the difference between the input signal and a multiple of 0.5 V. This difference can never exceed 0.5 V over the f.s.d. of the meter. Resistor R4 must therefore have a value that causes a f.s.d. at 0.5 V. The measured value is determined by adding the meter reading to the multiple of 0.5 V indicated by the LEDs.

The circuit is calibrated by first setting the meter to zero reading by P2. Next, apply a voltage equal to one fifth of the f.s.d. (here, 40 mV) to the input. Then, set P3 for minimum resistance, when none of the LEDs should light. Next, set P1 for f.s.d. Finally, adjust P3 until D7 begins to light and the meter reading falls to zero.

(R. Shankar)



030 AUDIO & HI-FI

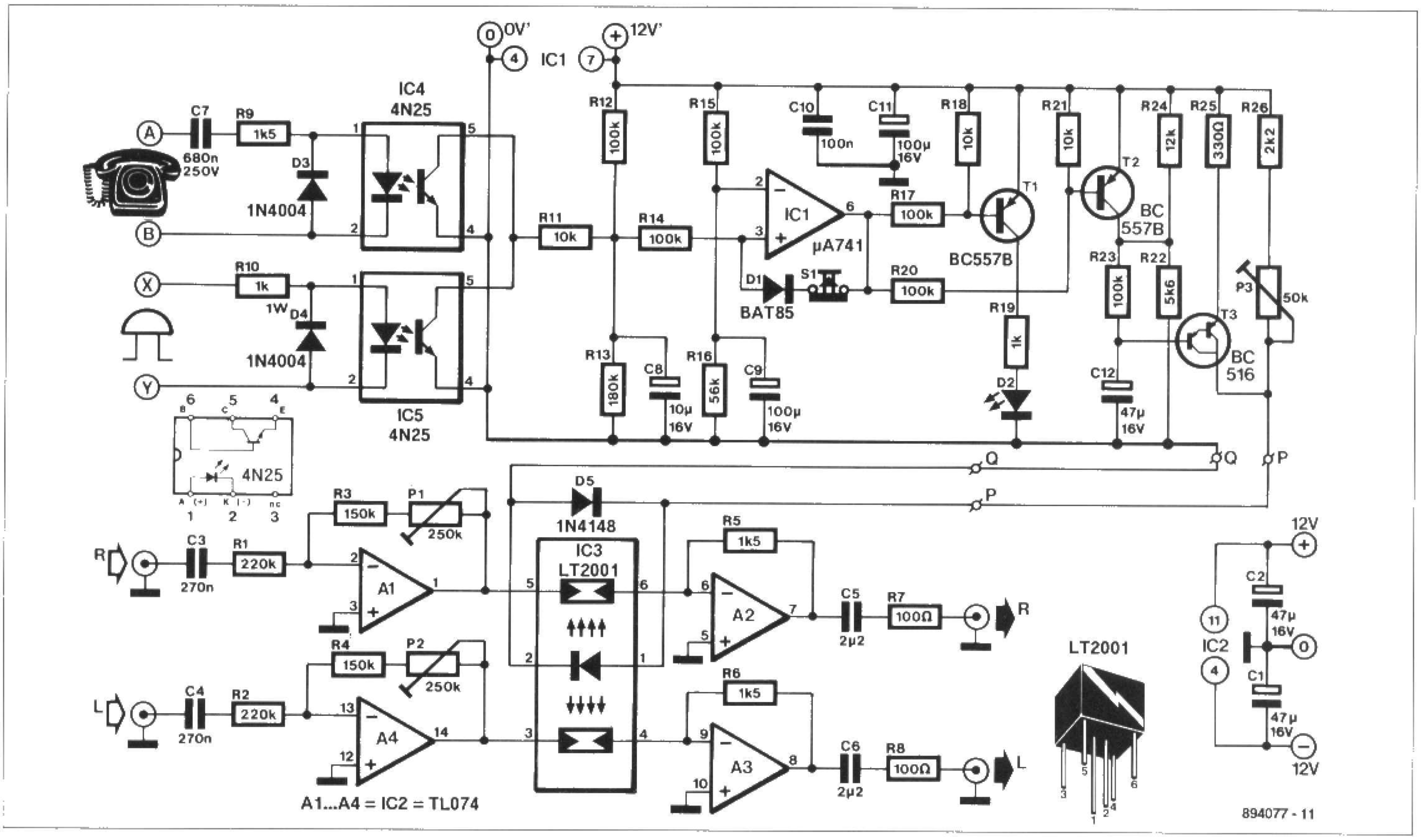
SOUND LEVEL ATTENUATOR

When the radio or record player is on at a fairly high volume, it is often impossible to hear the telephone or doorbell. A solution to this frequent difficulty is offered by this automatic attenuator. As soon as the doorbell or the telephone rings, it turns down the volume of the audio equipment.

The circuit consists of an optically controlled attenuator and the requisite electronics to connect it to, say, the telephone.

The attenuator is of fairly simple design and is based on a TL074. Its control part, consisting of a current-driven attenuator based on an LT2001 (a combination of an LED and two light-dependent resistors – LDRs – in a common enclosure), is incorporated in the audio equipment.

After the mains has been switched on, C9 causes the resetting of bistable IC1. The high voltage level at pin 6 of the 741 causes T1 and T2 to switch off: this, in turn, results in D2 not conducting and the voltage-controlled current source, T3, delivering maximum current (30 mA) to the LED incorporated in the LT2001. The illuminated LDRs will then have a value of about 1.5 kΩ. The voltage transfer of the attenuators can be preset (once and for all) to exactly 0 dB (at 1 kHz) by P1 and P2.



Terminals A and B of the circuit are connected direct to the corresponding terminals of the telephone (this may not be allowed in some countries – seek the advice of your local telephone manager), while terminals X and Y are connected across the doorbell terminals. Note that the doorbell must be fed from a 3–24 V transformer. If the telephone or the doorbell rings, the bistable will be set via the relevant opto-isolator (IC4 or IC5).

The low voltage level at the output of IC1 will cause T1 and T2 to switch on. This in turn causes D2 to light and C12 to charge via R23. Owing to the rising potential across the capacitor, the output of the current source will slowly diminish until the minimum value, set by P3, is reached. This has the effect that the TL074 turns down the volume click-free until a reasonable sound level, determined by the setting of P3, is reached.

Pressing switch S1 resets the bistable. This will cause D2 to go out, while the attenuation slowly drops to 0 dB. The attenuator is connected to the control electronics by two wires, P and Q. Thanks to the current drive, this (non-shielded) link may be up to 23 metres (75 ft) long.

The attenuator draws a current of only 10 mA and must be fed from a symmetric ±12 V supply, which may be taken from the audio amplifier. The control circuit needs an asymmetric +12 V supply and draws a current of about 35 mA.

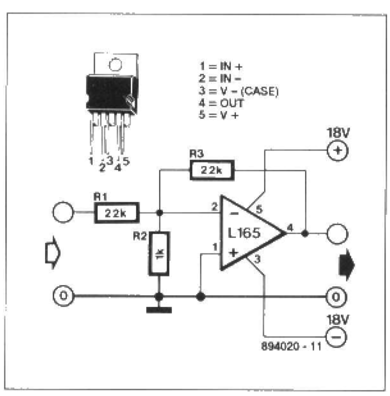
If the LT2001 is difficult to obtain, discrete components may be used: these should, of course, be fitted in a light-tight enclosure.

031 COMPONENTS

FAST UNITY GAIN OPAMP

A number of operational amplifiers can be used only in circuits that have a certain minimum amplification, because they have been designed with small internal compensation. The advantage of that is that the amplifier is faster.

If we look at two popular opamps, the LF356 and LF357, these characteristics are well illustrated. The LF356 may be used as a unity-gain amplifier with a gain-bandwidth product of 5 MHz and a slew rate of 12 V/µs. The LF357 needs an amplification of not less than 5 and has a gain-bandwidth product of 20 MHz and a slew rate of 50 V/µs.



It is possible to use the LF357 (and similar opamps) with smaller amplifications by using external compensation, yet retaining most of the bandwidth. Normally, this is achieved with a capacitor, but that is not the only, and certainly not the best, method.

An alternative method is illustrated in the diagram. Consider the L165 as a summing amplifier of which one input is connected to earth. It is clear that the second input then forms the input of a unity-gain amplifier whose amplification is determined by R1 and R3 (R3/R1 = 1). The unused input would have provided an amplification of 22 (R3/R2 = 22), which is rather more than the permissible minimum amplification of the L165. The opamp 'believes' that the amplification is higher than required. This has the benefit that the circuit has no tendency whatsoever to oscillate.

The ratio R3:R2 is only of value if R1 is very much larger than R2. Otherwise the amplification is R3/(R1//R2).

Note that the L165 must be fitted on a good heat sink, in spite of its internal thermal overload protection (Iout max. = 3 A).

032 COMPUTERS

PRINTER RESET

When during a computer print-out something goes wrong with the printer, such as the paper getting snarled up, the only way to stop the operation is normally to switch the machine off. That may be a useful, but certainly not an elegant, method: a reset knob, on the other hand, is.

Nearly all printers with a Centronics interface have a reset input at pin 31 of the Centronics connector (consult the handbook). That input is used in many MS-DOS systems to set the printer to a defined starting state and at the same time to empty the buffer.

The input may, of course, also be used to connect a reset switch to. The diagram in Fig. 1a shows how such a switch may be made quite easily. The 1 kΩ resistor prevents the computer output being short-circuited when the printer is being reset.

Users of the recently published printer buffer (Ref. 1) can fit the switch in the buffer or expand the existing switch so that the printer is reset at the same time as the printer buffer. The circuit is connected to the unused contact of S1 in the printer buffer. The existing reset switch may be expanded as shown in Fig. 1b.

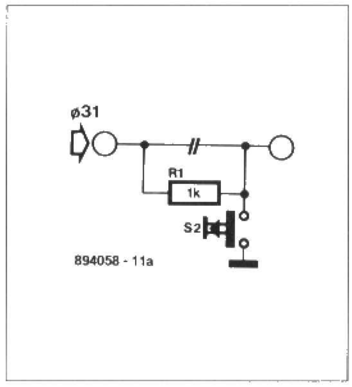


Fig. 1a

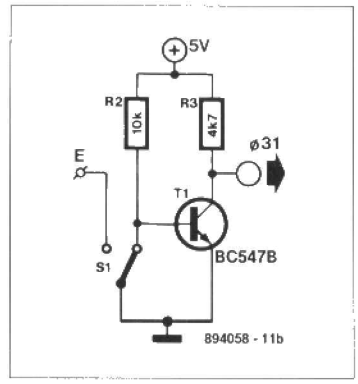


Fig. 1b

(1) "Centronics-compatible printer buffer", *Elektor Electronics*, March 1989, p. 21.

033 ELECTROPHONICS

VARIABLE LOW-PASS FILTER

The Type SSM2045 IC from PMI is an active low-pass filter whose order, Q-factor, cut-off frequency and amplification are set with the aid of control signals.

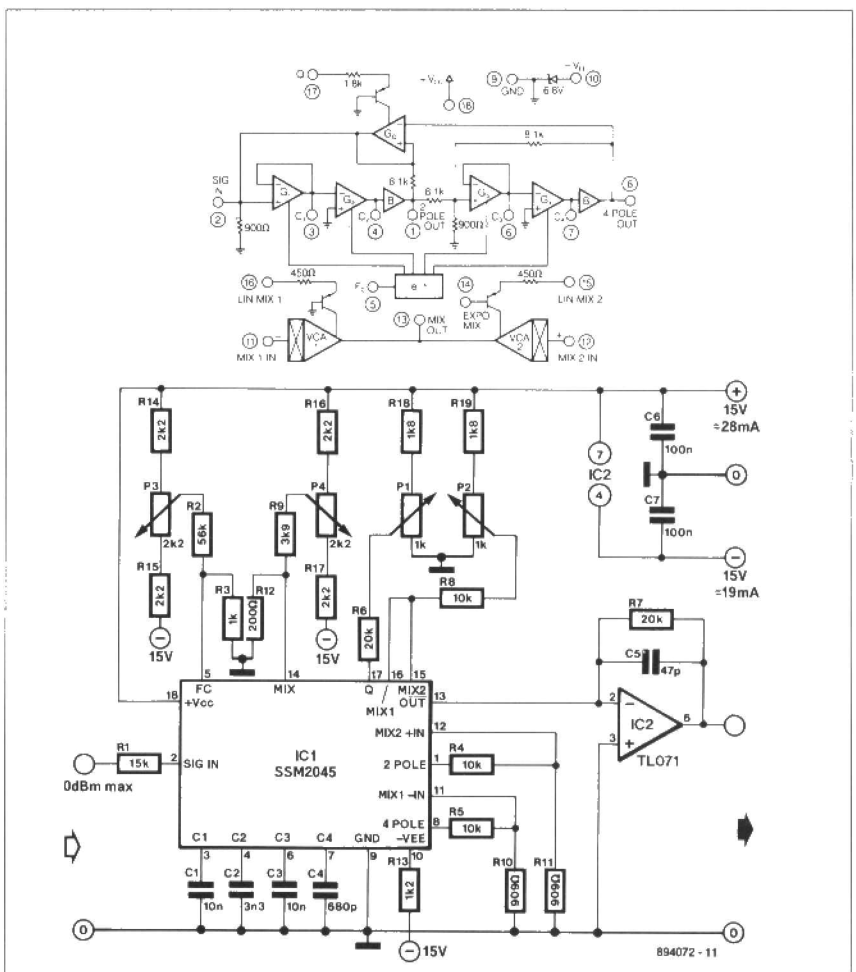
A possible design for use in electronic music systems is shown in the diagram. To prevent distortion, the level of the input signal is reduced by R1 and a resistor on board the chip to one not exceeding 150 V peak to peak.

Outputs 2 POLE and 4 POLE are each connected to an internal voltage-controlled amplifier (VCA), MIX1 and MIX2 respectively. For optimum off-set and control rejection, these connections are made via resistors R4 and R5. The gain of the VCAs is set by P2, which controls the current that flows through the amplifiers via pins 15 and 16. The maximum current at these control inputs is 250 μA. The balance between the VCAs, and with it the order of the filter, is set by P4 via pin 14. The voltage at this control input can be varied between –250 mV and +250 mV. The input must be driven from an impedance not exceeding 200 Ω. At a drive voltage of 0 mV, the VCAs attenuate the signal by about 6 dB.

The Q-factor depends on the current flowing into pin 17, which is controlled by P1. The input is protected by an internal 18 kΩ resistor. The Q-factor may be set so high as to cause the circuit to start oscillating. This happens when the current is between 120 and 185 μA.

The cut-off frequency can be shifted between 20 Hz and 20 kHz by varying the control voltage at pin 5 between +90 mV and –90 mV by P3. This voltage also determines the frequency of oscillation, enabling a variable oscillator to be created. The resulting sine wave has a distortion of about 1%.

The values of C1–C4 have been chosen to give the



filter a Butterworth characteristic if the current into pin 17 (Q) is zero.

The supply to the IC may be –5 V connected direct to pin 10, or up to –15 V via series resistor R13, through which a current of about 7.1 mA flows. The supply, –Vee, is limited internally by a 6.8 V zener diode.

The output current of the chip is converted into a voltage by IC2. The output of this stage has a small off-set voltage. If this can not be tolerated by the following equipment, the output must be taken via a coupling capacitor.

In the proposed circuit, the values of resistors R2, R6, R8, R9 and R12 have been chosen to allow control of the IC from voltages of 0–5 V or ±5 V.

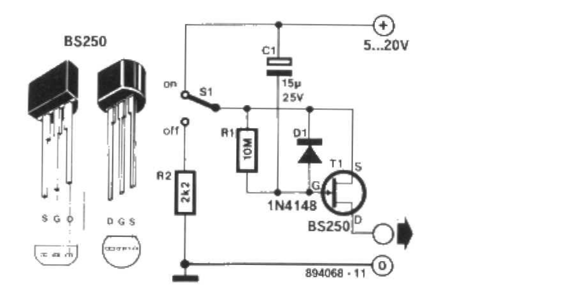
At an input signal of 0 dBm, the distortion is about 1%, which drops to 0.3% at –6 dBm and 0.03% at –20 dBm. The signal-to-noise ratio is of the order of 80 dB.

034 GENERAL INTEREST

AUTOMATIC SWITCH

Many battery-operated instruments, such as digital voltmeters, have a change-over on-off switch. The automatic switch presented here makes full use of that. When the instrument is off, C1 charges fairly quickly via R2 and D1. As soon as the instrument is switched on, C1 discharges slowly via R1. As long as the discharge current exceeds a certain level, T1 is switched on by the voltage across R1 and the supply to the instrument is on. When, after a few minutes, C1 is almost completely discharged, T1 toggles and the supply to the instrument is switched off. The period between switch-on and switch-off may, of course, be varied by different values of C1.

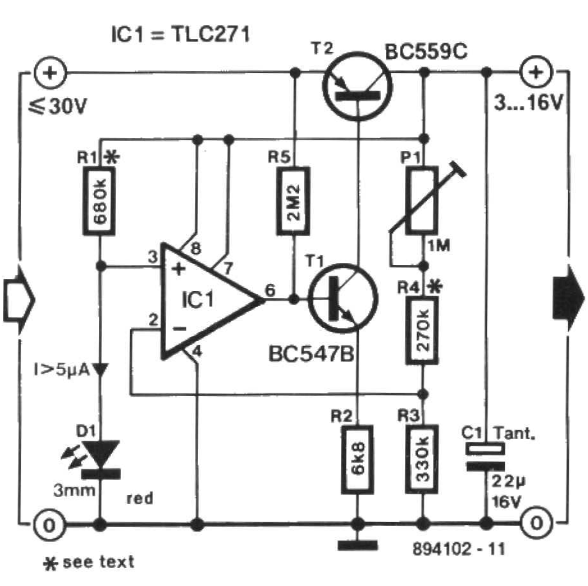
(Ph. Bosma)



035 POWER SUPPLIES

LOW DISSIPATION REGULATOR

With the advent of the now well-known three-pin voltage regulators, power supplies are taken for granted. Yet, there are cases where such a regulator is not wholly satisfactory. This type of regulator needs a fairly large potential drop across it (typically not less than 3 V) and draws a relatively high quiescent current (typically 6 mA for a 78xx). The regulator presented here is particularly attractive for battery-operated equipments and offers:



- variable, very stable output voltage;
- low potential drop (some tenths of a volt);
- very small quiescent current (20–30 µA).

In principle, the regulator is a normal series type. The voltage reference is obtained from a common or garden red LED that must draw not less, nor much more, than 5 µA. Even at that low current, an LED has a fairly stable voltage drop. To improve that stability, the current is drawn from the regulated output via R1.

Regulation is provided by CMOS opamp Type TLC271. This amplifier operates in the low bias mode, which ensures very low current consumption, by connecting pin 8 to the positive output terminal. The output of the opamp is used as base drive for series regulator T2 via current source T1. This configuration enables good control for only a small voltage swing at the output of the opamp. This is necesary since the slew rate of the opamp in the low-bias mode is pretty poor. The supply for the opamp is also taken from the regulator output. Capacitor C1 therefore serves as decoupling element for the opamp.

To obtain reliable control, a kind of bootstrap resistor, R5, was found necessary.

The values of R1 and R4 as shown in the diagram provide a variable output voltage of 3–8 V. Higher output voltages, up to a maximum of 16 V, are obtained by increasing R4 by 200 kΩ/V. Resistor R1 should also be increased in value, as long as the current through D1 does not drop below 5 µA.

In this type of circuit, great care should be taken to avoid parasitic capacitances resulting from long connections. These would cause a deterioration of the regulation.

The maximum output current depends mainly on the permissible dissipation in T2 and, therefore, to some extent on the difference between the input and output voltage.

036 TELECOMMUNICATIONS

DUPLEX AUDIO LINK

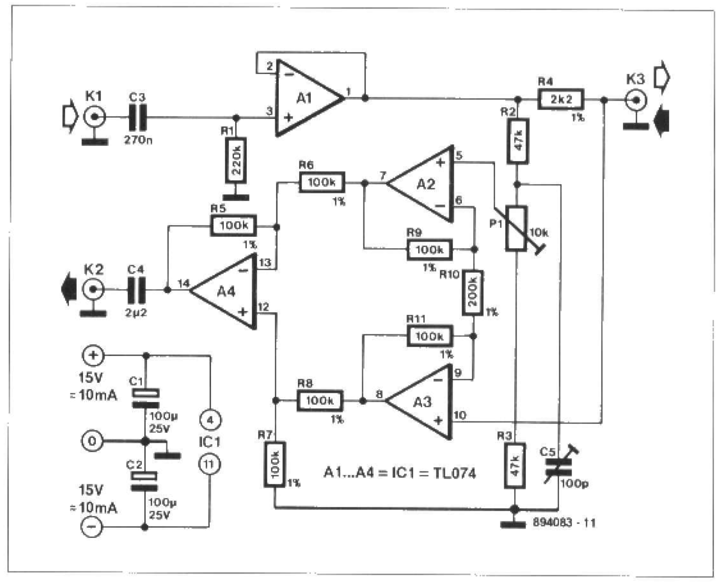
Duplex communication is, of course, not a new technique: it has been used, for instance, in telephone systems for many years. Those systems, however, make use of transformers to achieve duplex – the circuit presented here does it with the aid of electronics. The principle is fairly simple. Two senders impose signals U1 and U2 respectively on to the audio cable. The voltage across the cable is then (U1+U2)/2. The receivers at both sides of the cable deduct their side's sender signal from the cable signal: the result is the signal sent from the other end of the cable. This principle is the basis of the circuit shown. Note that a circuit like that is required at either end of the link.

Opamp A1 is connected as a buffer amplifier and serves as sender. The send signal is imposed on to the cable via R4. Terminating the cable by R4 results in the voltage across the cable being only half the voltage output of A1. This does not detract from the operation of the circuit, however. At the same time, R4 ensures that signals emanating from the other end of the link can not get to the output of A1; if they could, they would be short-circuited by the output.

The receiver is a differential amplifier consisting of opamps A2–A4. The quality of the differencial amplifier depends largely on the resistors used in association with the opamps and 1% types are, therefore, essential.

The cable signal, (U1+U2)/2, is applied to one input of the differential amplifier and the (halved) output signal of A1 to the other. Since the differential amplifier has a gain of 6 dB, the received signal applied to K2 has the same level as the original signal.

In practice, the proposed duplex system is not perfect and it is for that reason that, for instance, remnants of the sent signal are detectable in the receiver. Fortunately, these can be removed with the aid of P1. Furthermore, the cable used will load the output slightly capacitively, which causes the compensation voltage at the wiper of P1 to be not wholly in phase with the sent signal.



This effect may be virtually removed with the aid of C1.

The circuit is calibrated by connecting the cable to it and to its twin circuit and injecting a sinusoidal signal at a frequency of 1 kHz and a level of 5 V rms to its input. The input bus of the other circuit must be short-circuited during the calibration. Adjust P2 for minimum signal at K2. Next, increase the frequency of the input signal to 10 kHz and adjust C5 for minimum signal at K2. Repeat the procedure with the other circuit.

The signal suppression at 1 kHz is of the order of 80 dB, while at 20 kHz it is around 60 dB. These are pretty good values for this kind of circuit.

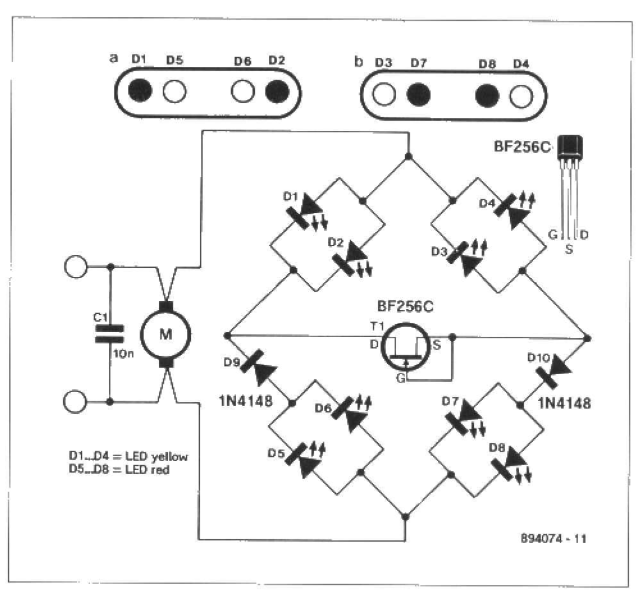
037 GENERAL INTEREST

HEAD / TAIL LIGHTS FOR MODEL RAILWAY

The price of a model railway locomotive is directly proportional to its facilities and finish. There are quite a few inexpensive ones on the market of which the finish leaves a lot to be desired. As as rule, manufacturers start their economy with the lights. The circuit presented here enables a DC locomotive to be provided with direction-independent head and tail lights. Since it uses LEDs, a very long life is guaranteed.

The circuit is based on a number of parallel-connected LEDs in a bridge network. The FET at the centre of the bridge ensures a constant current as long as the supply voltage exceeds 4.5 V. The brightness of the lights will, therefore, be independent of the supply voltage.

The LEDs are connected in parallel to keep the minimum operating voltage as low as possible. To ensure good current distribution, the pairs of parallel-connected LEDs should be of the same type and colour.



038 COMPUTERS

I/O-FRIENDLY KEYBOARD

Not all computers have a keyboard, yet it is often essential to have the use of one. Two circuits are presented here that enable a keyboard facility to be produced with the aid of only six or seven I/O lines.

Fig. 1 shows a circuit based on a 74HCT148 and a 74HCT138 that can serve 56 or 64 keys. The circuit in Fig. 2, based on a 74HCT147 and a 74HCT138, can address 72 keys via seven I/O lines. The choice between the two circuits depends on the number of available I/O lines and the wanted number of keys.

In either circuit, the key rows are selected by the bits on the A, B or C input of the HCT138. The combination of these bits determines which of the outputs Y0–Y7 goes low. As long as no key is depressed, the inputs of the HCT148 in Fig. 1 or the HCT147 in Fig. 2 are high. When a key is pressed, the inverted binary information at the output of the ICs show which key it is.

In Fig. 1, the $\overline{0}$ input of the HCT148 is not used, because the code associated with that input is the same as that generated when no key is pressed. The output, pin 14, of this IC is used to detect whether a key has been pressed. It goes low when a key has been pressed.

In Fig. 2, four 1s at the output indicate that no key has been pressed.

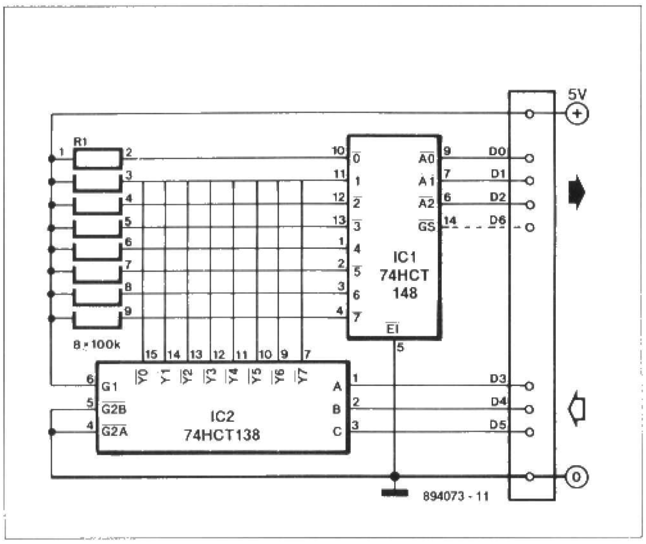


Fig. 1.

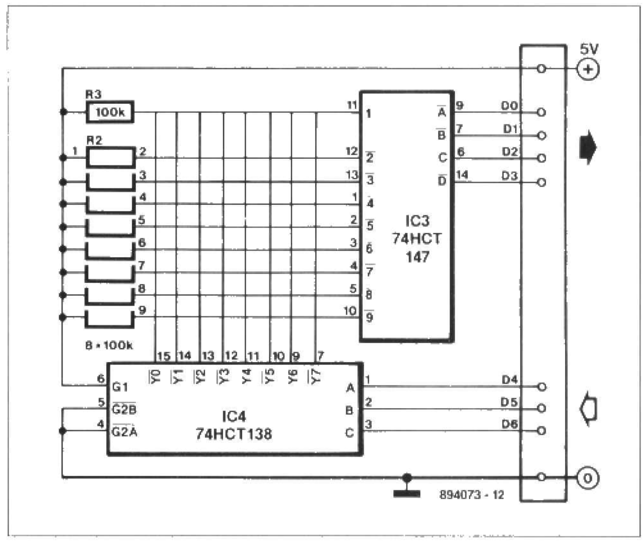
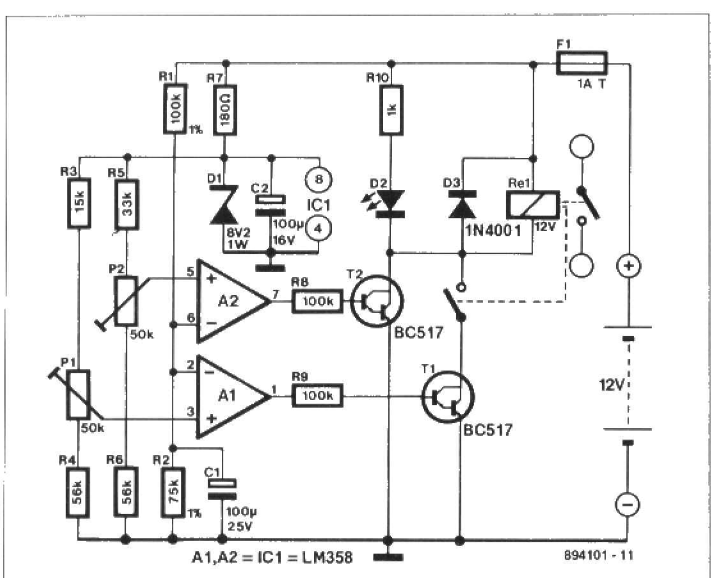


Fig. 2.

039 CAR ELECTRONICS

ENERGY CONTROL FOR BATTERY CHARGERS



In most automatic battery chargers, the power transformer remains connected to the mains even after the the battery (or batteries) has been charged. In many cases, considerable energy savings can be achieved by disconnecting the transformer from the mains when the battery is fully charged. This circuit performs this function for 12 V car battery chargers.

The battery voltage is monitored by an adjustable window comparator around opamps A1 and A2, which are powered by a stabilized supply voltage of 8.2 V (R7-D1). The high and low switching thresholds, U_H and U_L, are set by presets P1 and P2 respectively. The reference voltage for the opamp is obtained from junction R1-R2 and is a function of the battery voltage. With the given values of R1 and R2, a voltage divide factor, D, is obtained

$$D = R2 / [R1 + R2] = 0.43.$$

Taking into account the series resistors connected to the presets and the use of an 8.2 V supply voltage, the span of P1 is

$$7.2 / D = 16.7 \text{ V (max) to } 3.8 / D = 8.9 \text{ V (min)}$$

and that of P2 is

6.3 / D = 14.5 V (max) to 3.3 / D = 7.7 V (min)

In practice, it will be desirable to switch the charger off at a battery voltage of 14.0 V and on again when the voltage drops below 12.5 V, corresponding to a window of 1.5 V.

When the battery voltage is lower than U_L, it is, of course, also lower than U_H. This means that both T1 and T2 conduct, so that Re1 is energized. Contact K2 switches on the mains to the battery charger, and contact K1 keeps the relay energized even when T2 is switched off and the battery voltage rises to a value between U_L and U_H. When the battery voltage reaches U_H, both T1 and T2 are turned off, so that the relay is de-energized. After a while, however, it will be switched on again because the battery voltage will have dropped a little owing to the drop across its internal resistance.

Assuming that the required switching levels are U_L = 12.5 V and U_H = 14.0 V, the presets are adjusted as follows. Disconnect C1 and set P1 for U_H (max) and P2 for U_L (min). Power the circuit from a regulated supply set to 12.5 V and adjust P2 until the relay is just energized. Then increase the supply voltage to 14.0 V and adjust P1 until the relay is just de-energized. Finally, connect C1 again and connect the circuit to the battery terminals.

The relay must be rated at 12 V DC and 300 Ω. It must have two make or change-over contacts, of which at least one has a voltage rating higher than the mains voltage. One suggested type is the V23037-A2-A101 from Siemens.

The circuit draws 25 mA, which rises to 65 mA when the relay is energized.

(M.S. Dhingra)

040 TEST & MEASUREMENT

TTL SUPPLY MONITOR

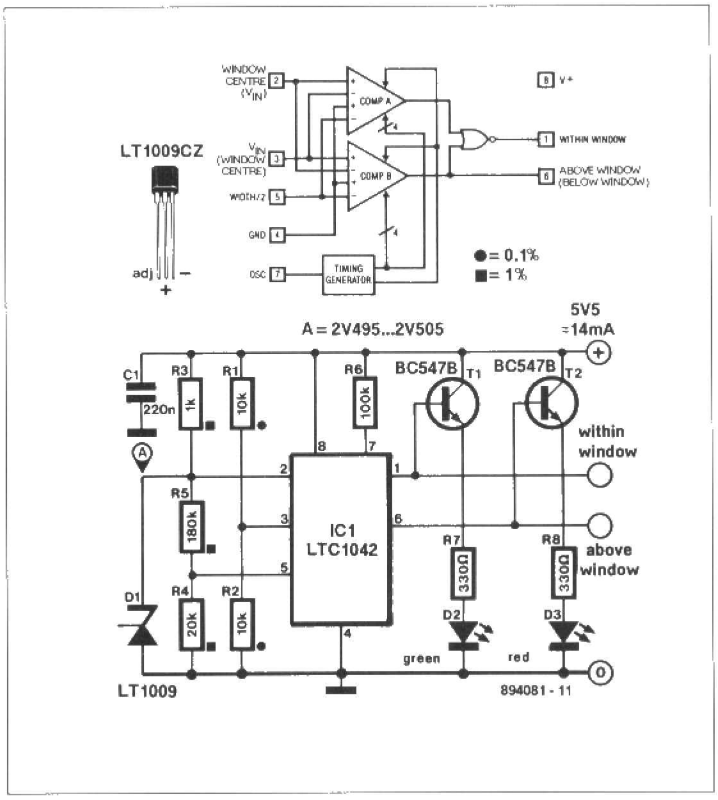
The Type LTC1042 IC from Linear Technology is a window comparator that can function with very small currents. This is made possible by the use of sampling techniques that enable the disabling of certain parts of the IC in the non-active phase. This economical behaviour is not so important in the present circuit: the monitor draws a current of rather more than the minimum 100 µA.

The comparator is set with the aid of a bandgap reference diode, D1. The reference voltage of 2.5 V provided by this diode is connected direct to the window centre pin 2.

The width of the window is also fixed with the aid of the reference voltage. Since the circuit is intended to monitor a TTL supply (5 V), the width of the window is arranged at one tenth of this, which is a convenient sub-multiple of the reference voltage. Potential divider R4-R5 gives a voltage of 0.25 V at half-window-width pin 5. This arrangement causes the within window output to go high when the input voltage (Vin at pin 3) exceeds 2.5 V ±10%. The input voltage is held at exactly half the supply voltage by potential divider R1-R2.

Transistors T1 and T2 drive indicator LEDs. When D2 lights, the circuit and the supply voltage are all right. When D3 lights, the supply voltage is too high. If neither of the diodes lights, the supply voltage is too low or even absent.

If you want a LED to indicate that the supply voltage is too low, interchange the function of pins 2 and 3. The above window output then becomes a below window output. Note that the circuit then needs a separate power supply, otherwise the below window LED can not be actuated.

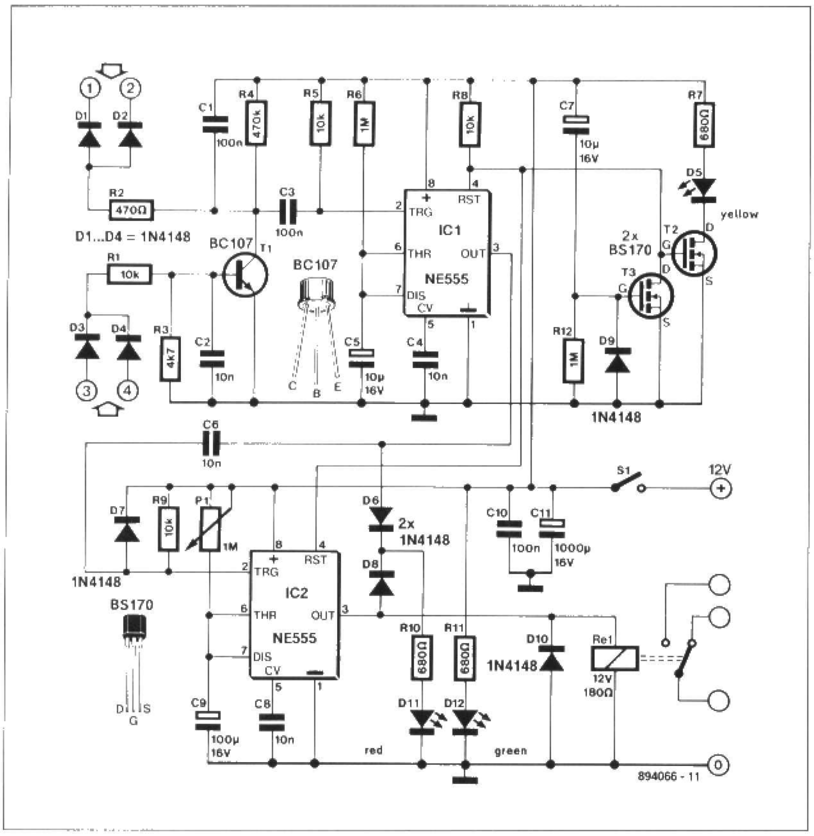


041 CAR ELECTRONICS

CAR ALARM

The car alarm accepts signals from a variety of sources, including special sensors and the standard switches in a car, such as the door and ignition contacts. The unit has a relay output that enables controlling an acoustic transponder (loudspeaker, buzzer), light, a radio transmitter, and others.

The alarm is remarkably simple to set, because all control functions are performed by a single switch. It is switched on after parking the car by closing S1 when the green LED lights to indicate that the driver has 13 seconds to leave the car. When this delay has lapsed, a yellow LED lights to indicate that the alarm is set. When any of the alarm sensors is actuated (which also happens when the rightful owner of the car opens a door), a red LED lights.



The relay will then be energized after 17 seconds unless switch S1 is opened in the mean time. Since the location of that switch is known only to the car owner, anyone entering the car illegally will set off the alarm.

Two timers Type NE555 are connected as monostable multivibrators. Transistor T1 and associated components form the triggering stage, which is sensitive to both positive and negative triggering signals supplied by the sensors or switches. MOSFETs T2 and T3 serve to provide the 'delayed disable' function and to drive the 'alert' LED respectively.

When the unit is powered, D12 lights, but the alarm remains off for the time defined by time constant R12-C7. Since the capacitor is initially discharged, T3 conducts and pulls the reset pins of both timers to ground until the voltage across R12 has dropped to about 3 V. That level enables both timers, and causes the 'alert' LED to light. When the power is switched off for a second and then on again, C7 is dicharged rapidly via D9: this sequence resets the alarm to its initial state.

When enabled, the alarm may be actuated by a low level at input pinss 11 or 12 (door switches) or a high level applied to inputs 13 or 14 (ignition system power). The number of inputs may be expanded as required by adding one diode for each sensor or switch. When only one of the inputs is used, it is still recommended to use a diode.

Any trigger pulse or change in the direct voltage at the inputs triggers timer IC1 via C3 and C5. The output pin, 3, of IC1 is driven high so that the 'actuated' LED (D11) lights. Diodes D6 and D8 form an OR function to decouple the outputs of the timers. Capacitor C6 is discharged in preparation of the triggering of IC2. After the delay set by R6-C5, the output of IC1 goes low, thereby triggering IC2 via R9 and C6. The output relay is energized and the transponder is powered by the car battery via the relay contact. The alarm sounds for about 65 seconds (defined by R9-C10). The unit may be reset at any time, however, by operating S1 for at least a second.

(R. Lalic)

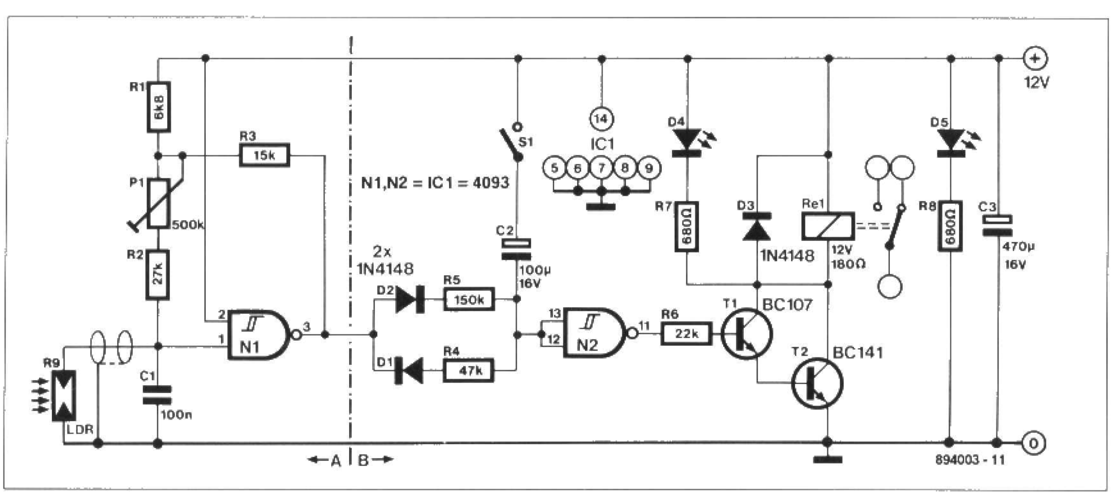
042 GENERAL INTEREST

TWILIGHT SWITCH

This inexpensive unit switches the light on at dusk and off again at dawn. The circuit has separate time bases for the on and off delays. The dotted line in the diagram divides the circuit into two halves: A, the light-dependent switch around gate N1; and B, the on-off time base around N2 and LED and relay driver T1-T2.

The voltage at junction R2-R9-C1 is inversely proportional to the light intensity measured by light-dependent resistor (LDR) R9. Schmitt trigger gate N1 toggles whenever this voltage reaches one of the input threshold levels. Since the difference between these levels is large compared with the voltage span produced by the potential divider, a variable feedback loop is provided to achieve an effective switching span of 300 to 400 mV. When the output of N1 is high, the voltage at junction R1-P1 is almost equal to the supply voltage. When the output is low, the voltage drops to the level required for the threshold difference at the input of N1.

The output of N1 drives two time base circuits: C2-R4-D1 for the 'on' state and C2-R5-D2 for the 'off' state. These networks switch the output of N2 on and off after the wanted delays. The lamp relay and an indication LED are driven by darlington stage T1-T2, which is controlled by the output of N2.



Capacitor C1 prevents HF signals picked up by the cable between the LDR and the switching unit causing spurious triggering. Because of the high output impedance of the Type 4093, the cable should be a screened type.

The remaining two gates in

the 4093 package may be used for duplicating the time base (part B) to obtain a switching sequence. The values shown for R5-C2 and R4-C2 form a good starting point for dimensioning the delays introduced in this additional time base. The lowest permissible value of R4 and R5 is 47 kΩ: their highest value depends mainly on the leakage current of C2.

During the testing of prototypes of this circuit it was noted that the switching threshold and hysteresis depend on the make of the 4093. Good results were obtained with SGS's HCF4093BE; devices from other manufacturers may require the value of R3 to be slightly different. The inputs of the unused gates in the 4093 must be earthed.

The LDR is mounted in a suitable waterproof housing, screened from direct light sources.

Preset P1 serves to adjust the light intensity level at which the circuit switches. To prevent too slow a response to changes in the setting, the timer should be aligned with S1 open.

The current drawn by the timer is mainly that drawn by the energized relay.

(R. Lalic)

043 CAR ELECTRONICS

PSYCHOLOGICAL CAR LOCK

The lock circuit is based on elementary psychology rather than on any recent development in electronics.

The lock consists of a 12-key membrane keypad and an associated visual indication circuit. The complete unit is installed in an out-of-the-way position below the dashboard in a car, where it is not too difficult for a potential car thief to spot.

When the keys on the keypad are pressed, the impression is given that the lock will enable the car ignition when all four LEDs light. Because of the special configuration of the lock circuit, this will never happen. Eventually, the would-be thief gets frustrated (we hope) and tries another car, not realizing that the lock circuit is simply not connected to the ignition circuit. Only the rightful owner knows that the car can be started after a special switch, marked, say, 'wiper' is operated. This switch, installed as an accessory on the dashboard, is connected in series with the positive supply line to the iginition coil.

The membrane board must be a type with a common line, not one with a matrix configuration. A suitable keyboard may, of course, also be made from individual keys with numbered caps. The lines marked K0–K9 in the diagram must go to the associated key number on the keypad. Keys ✻ and # are non-connected dummies. The four LEDs are fitted in a row near the keypad, and give an indication that suggests that the combination entered is correct.

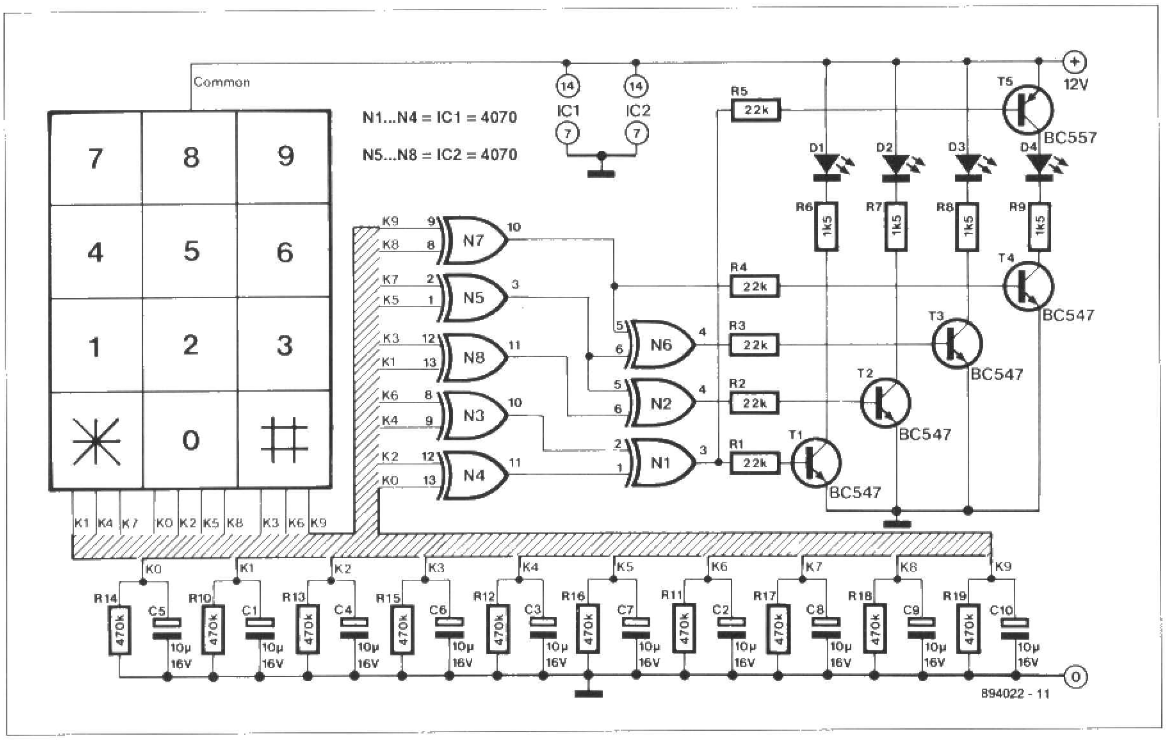
As already stated, the circuit makes it impossible for all four LEDs to light simultaneously. This is because pin 3 of N1 is logic high when D1 is on. This means that T5 is off, so that D4 can not light in spite of all other key combinations. It also means that D4 can not light unless D1 is out.

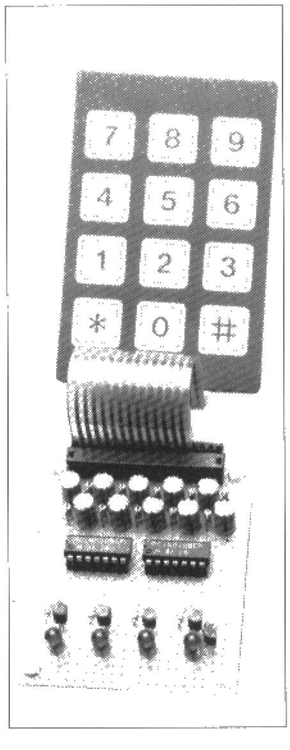
Eight XOR gates in two 4070 packages determine which LEDs are on for a particular code entered via the the keyboard. RC networks connected to each keyboard line keep this active for about four seconds after the key has been released. When, for instance, key 1 is pressed briefly, the voltage on line K1 rises to the supply level. Since pin 13 of gate N8 is then the only logic high input of all XOR gates, pin 4 of N2 goes high. This causes T2 to conduct and D2 to light. After about 4 s, the voltage on C1 has dropped to a level that N8 recognizes as logic low, and D2 goes out. Within the 4 s period it will, however, go out the instant key 5 or 7 is pressed. Diode D3 then lights, followed by D2, which remains on for a short period. The two LEDs then go out simultaneously. Note that this functional description applies to only one of many possible combinations.

Apart from its use in cars, the circuit may also find application in games.

The current drawn by the circuit is mainly that drawn by the LEDs that are on during the 4 s interval. In stand-by mode, less than 1 mA is drawn.

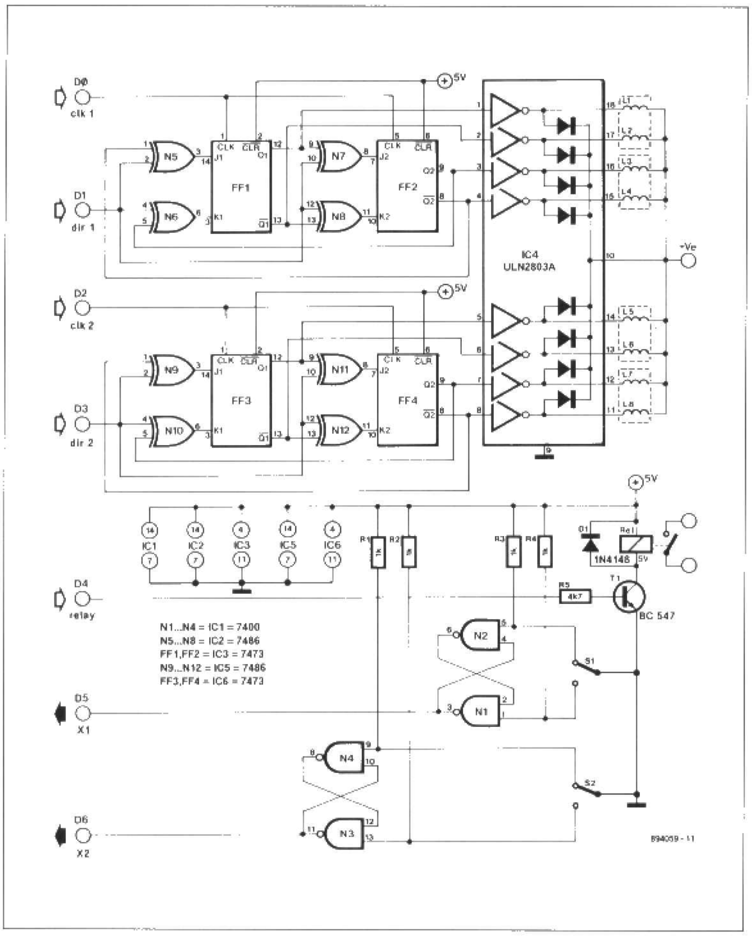
(C. Sanjay)





044 GENERAL INTEREST

X–Y PLOTTER INTERFACE



This low-cost circuit can drive two stepper motors and a relay by digital control data supplied from a computer. It can also detect the position of two microswitches and supplies logic levels back to the computer as positional information. This combination makes the interface ideal for use as an *X–Y* plotter or for building a buggy-style robot.

Circuit IC1 is configured as two set-reset (S-R) latches to provide contact debouncing for the two microswitches, S1 and S2, whose position is detected by computer reading port bits D5 and D6. The relay drive circuit around T1 may be used to switch a solenoid-operated pen on and off under the control of port bit 4.

Motor drive is provided by XOR gates N5–N12, bistables FF1–FF4 and integrated motor coil driver IC3. This combination can drive two unipolar 4-phase stepper motors in half-step mode by setting the direction of rotation with the aid of bits D1 and D3, and a high-to-low logic pulse transition on bit D0 or D2. The control functions are summarized in the table. The motor driver, IC4, is capable of sourcing 500 mA per phase at a maximum motor voltage of 50 V. The ULN2803A has internal diodes that protect it against reverse EMF generated by the motor coils when they are deactuated.

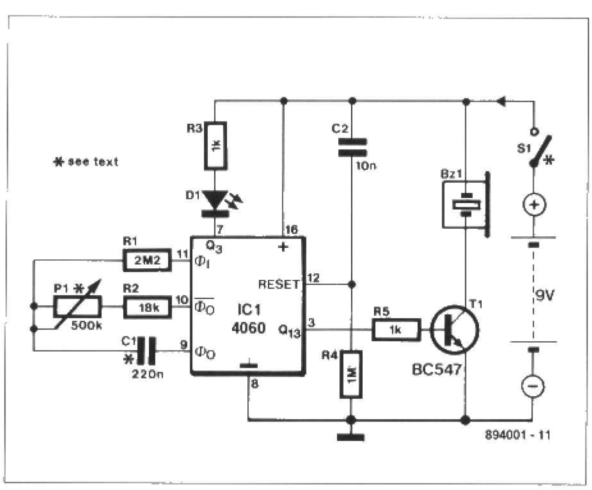
To drive the interface from a computer, set the I/O port for five output bits, D0–D4, and two input bits, D5 and D6, and send the appropriate control signals to the circuit.

The interface requires two supply voltages: 5 V for the logic circuits and the relay driver, and 12 V (+Ve) for the stepper motor coils. Motor 1 is L1–L4 and motor 2 is L5–L8.

Step pulse	Motor 1				Motor 2			
	L1	L2	L3	L4	L5	L6	L7	L8
0	1	0	1	0	1	0	1	0
1	1	0	0	1	0	1	1	0
2	0	1	0	1	0	1	0	1
3	0	1	1	0	1	0	0	1
0	1	0	1	0	1	0	1	0

045 GENERAL INTEREST

TIMER WITH AUDIBLE WARNING



Applications of this little circuit include a portable parking meter timer and egg timer. The 14-stage binary ripple counter Type 4060, IC1, has an on-chip oscillator capable of stable operation over a relatively wide frequency range. In the present circuit, the oscillator frequency is determined by an external RC network connected to pins 9, 10 and 11.

When the circuit is switched on with S1, the pulse at junction R4-C2 resets the counter and counting starts. When the count reaches bit 14 (Q13), pin 3 goes high so that the self-oscillating piezo-electric buzzer, a 12 V type, is turned on via driver T1.

The time delay is set with the aid of P1. Time delays of between one minute and two hours are possible by appropriate dimensioning of the timing components:

1–30 minutes: C1 = 220 nF; P1 = 500 kΩ
1–60 minutes: C1 = 470 nF; P1 = 500 kΩ
1–120 minutes: C1 = 470 nF; P1 = 1 MΩ

The timer is powered by a 9 V PP3 battery. Light-emitting diode D1 does not affect the operation of the circuit and is included merely to show that the timer works. Diode D1 and resistor R3 are, therefore, optional components. A mercury tilt switch may be used for S1 if the unit is to be used as a kitchen timer. The timer is then started by inverting it like a sand-glass.

With the buzzer actuated, the timer draws a current of about 10 mA.

(R.G. Evans)

046 TEST & MEASUREMENT

HCMOS SQUARE WAVE GENERATOR

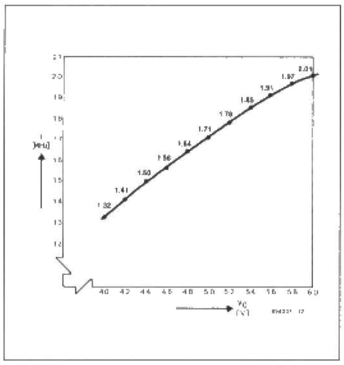
This pulse generator supplies three rectangular output signals with a duty factor of 0.5. The signals have a fixed phase relation to one another: output 3 is the reference; output 2 has a phase shift of about 180 °; and output 1 has a phase shift of about 10 °.

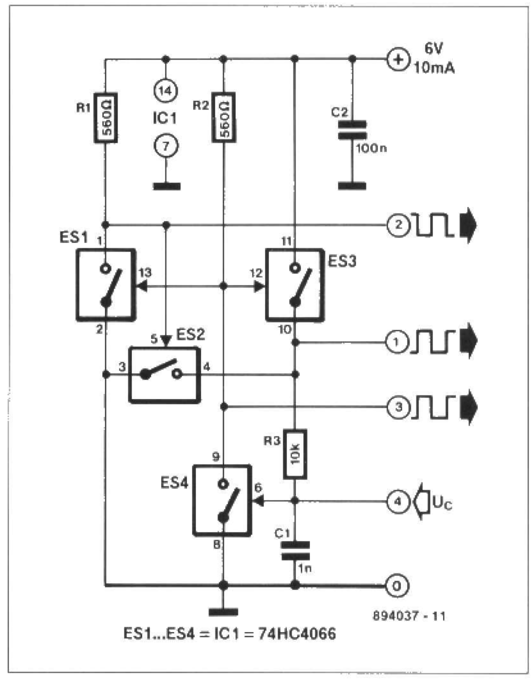
The generator is formed by the four bi-directional HCMOS electronic switches contained in a Type 74HC4066 IC. Its operation is based on the fairly accurately defined switching threshold of an HCMOS input. The toggle point for low and high levels lies around $Ub/2$ to ensure a duty factor of 0.5 and thus a square wave output signal.

When the supply is switched on, C1 is charged via R3 and the on-resistance of ES3. When the voltage of C1 reaches $Ub/2$, ES4 closes and pulls the control input of ES3 low, causing C1 to discharge via ES2 and R3. When the 'low' switching threshold is reached at the control input of ES4, the generator starts to oscillate.

Considering the oscillator is basically an RC type, its stability is pretty good.

The output frequency of the generator is a function of the control voltage, Uc, as shown by the accompanyhing characteristic.





047 POWER SUPPLIES

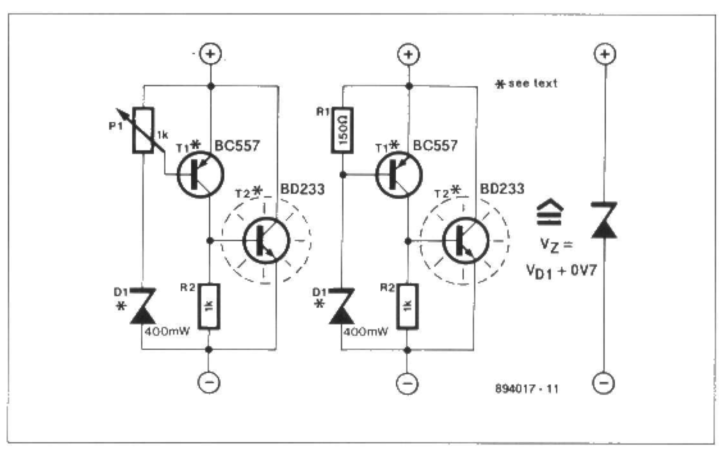
HIGH-POWER ZENER DIODE

Although its regulation characteristics are not as good as those of an integrated voltage regulator, a high-power zener diode has, nevertheless, useful applications, for instance, in shunt regulators.

The circuit shown in the diagram simulates a fairly expensive and hard-to-come-by high-power zener diode. Basically a two-stage current amplifier with a low-power (400 mW) zener reference, it is capable of sinking up to 500 mA at a maximum 25 V. The effective zener voltage of the circuit is about 0.7 V higher than that of the reference device, D1.

Using a preset, P1, instead of a fixed resistor, R1, enables the output voltage defined by D1 to be increased slightly. This substitution is useful when, say, a 6V2 or 9V1 zener is to be simulated and a low-power type of this rating is not available. The next lower value in the E6 range (here, 5V6 or 8V2) may then be used for D1.

When the voltage across the circuit increases beyond the zener voltage, T1 conducts and provides a base current to T2. This power transistor then passes virtually all the current the 'zener' is capable of sinking. The dimensions of the heat sink for T2 depend on the maximum expected dissipation.

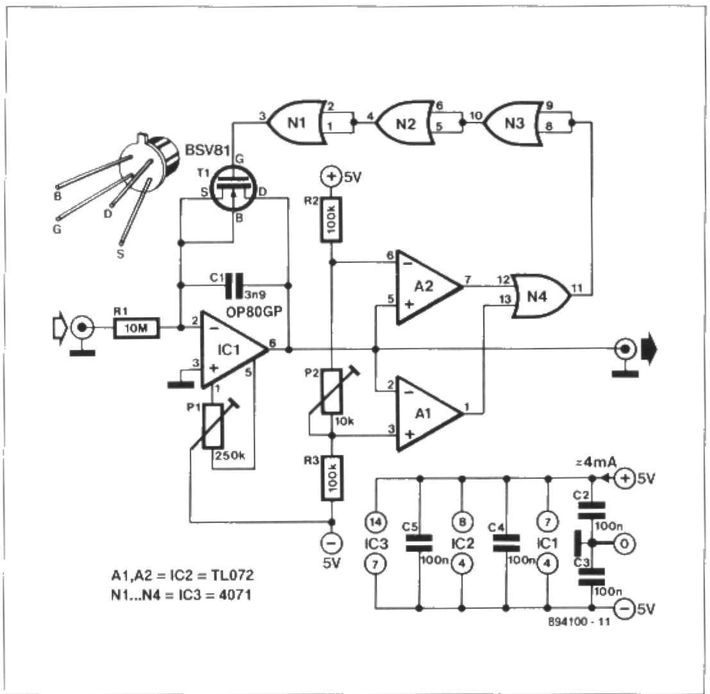


Transistor T1 should be a PNP type with a fairly high current gain, for instance a BC557B or BC559C, while T2 may be almost any medium or high-power NPN type, for example, a BD135 through BD139, BD241, TIP31 or 2N3055.

(C. Sanjay)

048 TEST & MEASUREMENT

VOLTAGE-CONTROLLED OSCILLATOR



The voltage-controlled oscillator (VCO) presented here is based on a Type OP80 operational amplifier. This opamp has an exceptionally low bias current of, typically, about 200 fA (femto-ampere = 10^{-15} amp) and 2 pA maximum, so that any off-set caused by this current is minute. It is, therefore, ideal for use as an integrator, because the operation of that kind of circuit is affected readily by off-sets.

The OP80-based integrator in the diagram is used as a VCO that is not affected by the polarity of the control voltage. A direct voltage at the input of the circuit will cause C1 to charge. Depending on the polarity of the input voltage, the potential across C1, and thus the output voltage of IC1, will be positive or negative. The speed with which C1 charges depends on the magnitude of the input voltage: this characteristic is used to generate a signal at a voltage-dependent frequency. To that end, the outp ut signal of IC1 is applied to a window comparator that has a switching threshold for both the positive and the negative maximum signal. These maxima are set to ±100 mV by P2. In some cases, it may be advantageous for symmetry to split R2 or R3 into a fixed resistor in series with a preset potentiometer.

When one of the comparators toggles, T1 is switched on via N1–N4 so that C1 discharges. This results in a neat sawtooth signal at the output of the circuit, whose frequency depends on the input voltage. Gate N4 ensures that the FET reacts to both comparators. The other three gates delay the switching signal slightly to ensure that the FET is switched on long enough to allow C1 to dicharge completely.

The Type BSV81 MOSFET is provided with a separate substrate connection that must be linked to the source. Since the substrate is already connected internally to the housing, the device is very sensitive to random radiation, so that the oscillator is best fitted in a small metal enclosure.

If a BSV81 is not obtainable, another MOSFET with very low R_{DSon} and very small C_{rs} may be used. If that also is not possible, a junction FET may be tried, but in that case a diode must be connected in series with the gate and a resistor of about 10 kΩ between the gate and the negative supply line. It is important to ensure that the pinch-off voltage level is reached readily. It may well be necessary to experiment with the value of C1.

Correct dimensioning of R1 will enable the relation between input voltage and frequency to be set at, say, 1 Hz/mV. With the input short-circuited, adjust P1 for the lowest possible frequency of the output signal (ideally, $f = 0$). The maximum input voltage is determined by the peak output current of IC2 (15 mA) and amounts to $15 \times 10^{-3} \times R1$.

The output signal of the VCO is a clean sawtooth signal at a frequency of up to 10 kHz, although higher frequencies are possible. The frequency as a function of the input current is given by:

$$f = I_{in} / (U_{top} \times C_1) \qquad [\mathrm{Hz}].$$

With values as shown in the diagram,

$$f = I_{in} / (3.9 \times 10^{-10}) \qquad [\mathrm{Hz}].$$

Finally, note that the supply voltage to the OP80 must under no circumstances exceed ±8 V. The circuit draws a current of typically 4 mA.

049 COMPUTERS

MONITORING TEMPERATURE WITH THE C64

Maplin's module Type FE33L provides an inexpensive and convenient means of monitoring temperature. The module has a built-in A-D converter and an LC display and works from a single 1.5 V battery. Since it is often impractical to take frequent readings manually, the module provides a serial data output that can be used with most microprocessor systems. The combination of hardware and software given in this article enables a C64 computer to use the serial data, within BASIC, via the USR function.

The hardware consists of nothing more than a simple TTL level driver and may be mounted on a small piece of prototyping board. This may be connected to the module by three short wires, while the outputs go to a two-by-twelve 0.156" pitch edge connector for the C64's user port. Pins 5 and 16 of the module should be short-circuited to obtain the maximum sampling rate of one per second. Check all connections before switching the computer on.

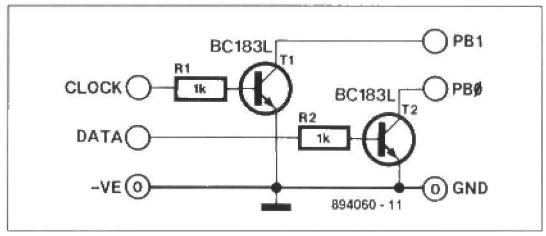
The listing provided loads a machine-code program into the small section of RAM above the BASIC ROM at location 49152 ($C000). Note that some lines are very similar to others thus assisting entry. Once this has been RUN (without errors), and SYS 49152 has been entered, the temperature is obtained as follows:

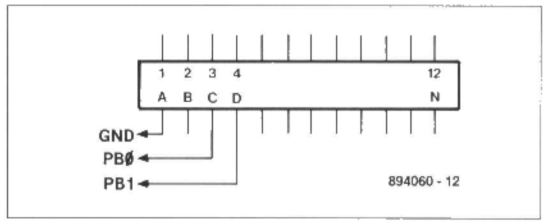
```
TEMP = USR(0) : PRINT TEMP
```

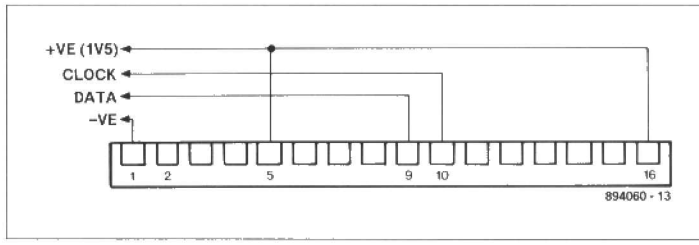
This line can be incorporated into any BASIC program.

```
0 REM *********************************
1 REM *** Maplin Temperature Module ***
2 REM ***      Interface Software   ***
3 REM *********************************
4 REM *** SYS 49152 to initialise   ***
5 REM *** Temperature = USR(0)      ***
9 :
10 FOR L = 0 TO 178:READ A:T=T+A
11 POKE 49152+L,A:NEXT L
12 IF T<>23290 THEN ? "Error in DATA "
13 :
14 DATA 162,0,142,3,221,169,16,141
15 DATA 17,3,169,192,141,18,3,96
16 DATA 120,169,2,162,0,44,1,221
17 DATA 208,251,172,1,221,232,44,1
18 DATA 221,240,250,224,90,48,236,162
19 DATA 13,152,41,1,157,179,192,202
20 DATA 169,2,44,1,221,208,251,173
21 DATA 1,221,41,1,157,179,192,202
22 DATA 169,2,44,1,221,240,251,224
23 DATA 0,208,229,162,13,189,179,192
24 DATA 73,255,41,1,157,179,192,202
25 DATA 208,243,160,0,32,162,179,32
26 DATA 12,188,162,12,160,4,10,24
27 DATA 125,179,192,202,136,208,247,32
28 DATA 155,192,10,24,125,179,192,202
29 DATA 136,208,247,32,155,192,10,24
30 DATA 125,179,192,202,136,208,247,168
31 DATA 32,162,179,32,106,184,32,254
32 DATA 186,173,192,192,240,3,32,180
33 DATA 191,88,96,142,191,192,168,32
34 DATA 162,179,32,106,184,32,226,186
35 DATA 32,12,188,169,0,160,4,174
36 DATA 191,192,96
```

The first part of the machine-code program sets the USR vector and all the port B lines to input, while the remaining code is called by the USR function itself. When called, the program waits for the primary clock pulse, which is longer than the others, and then reads in each subsequent bit from the data line. These bits are converted from BCD format into a single floating-point







that is returned by the USR function. The software will behave correctly only when the module is in the default °C mode, but this is not a restriction as readings can be converted readily to another scale. If the device is to be used for serious control applications, it must be borne in mind that the software will wait patiently for the primary clock signal to arrive from the module. If the clock signal fails for any reason (for instance, a break in the cable), the control program will be left hanging in an endless loop. It is, therefore, recommended to use a non-maskable interrupt (NMI) generated by the timers on CIA #2 to interrupt the program after a specific duration (for example, greater than the expected sampling time) and return some sort of error condition. For simple applications, this is not necessary and no further programming is required.

(J. Pelan)

050 POWER SUPPLIES

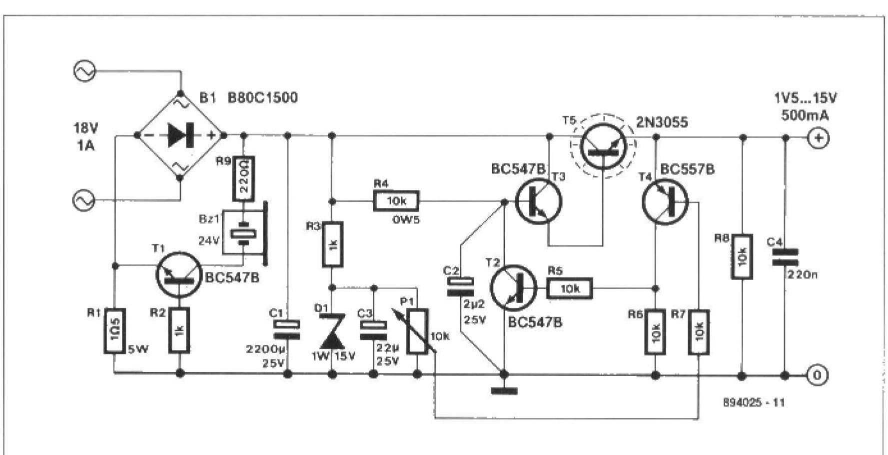
SIMPLE VARIABLE POWER SUPPLY

This low-cost power supply has an output voltage range of 1.5–15 V at a maximum current of 500 mA. Regulation is better than 2% for output currents not exceeding 350 mA. Voltage adjustment is effected by a potentiometer and an acoustic overload indication is provided.

Transistor T4 compares the voltage at the wiper of P1 with the output voltage. When this is 0.65 V higher than the set voltage, T2 is switched on, which removes the base current from darlington power stage T3-T5. In this manner, the output voltage of the supply is 0.65 V higher than the reference potential at the base of T4, which is derived from a 15 V zener diode, D5.

The voltage at the 18 V, 1 A winding of the external mains transformer is rectified by bridge B1 and smoothed by C1. A simple acoustic overload alarm (Bz1) is actuated when the output current exceeds around 500 mA. Note that the exact level of actuation depends on the electrical specification of the buzzer, which should be a 24 V, self-oscillating type.

The power supply is, in principle, not short-circuit proof, although the use of a generous heat sink for T5 will enable that transistor to withstand the maximum dissipation of about 20 W for the few seconds that lapse before the supply is switched off.



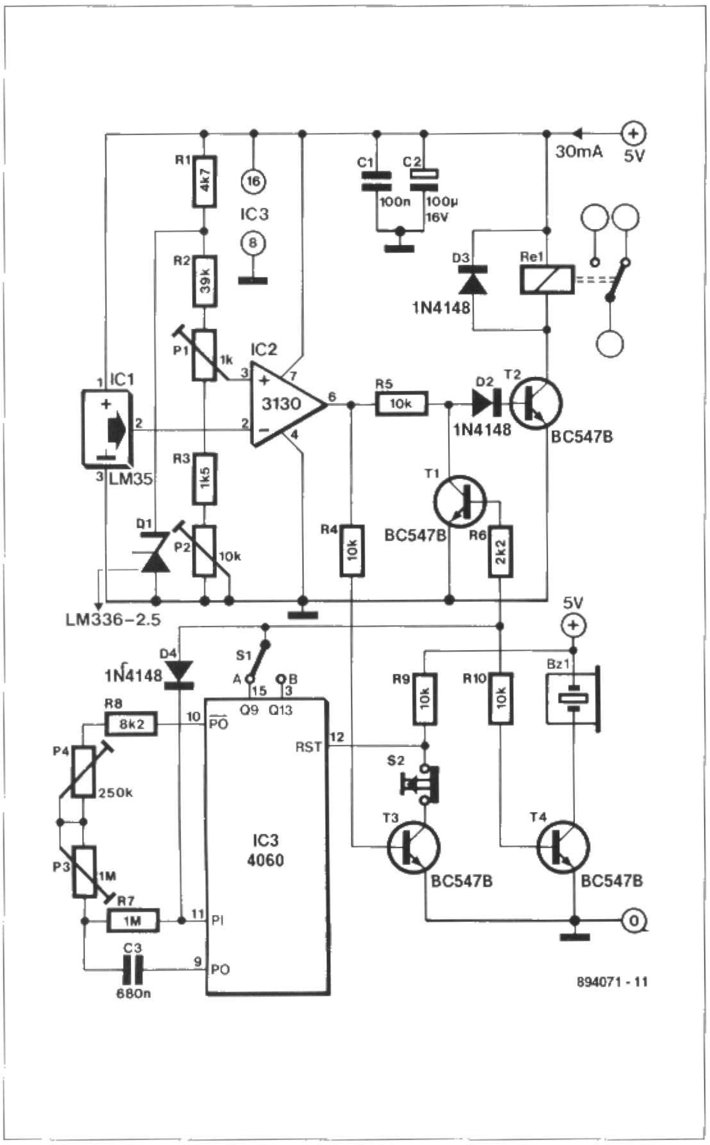
To obviate radiation, the supply must be fitted in a metal enclosure. Interconnections should be kept as short as possible.

Capacitors C2 and C3 should be tantalum types.

(P. Sicherman)

051 GENERAL INTEREST

HEATING TIMER



This timer has temperature and time settings. The temperature range is about 150 °C and the time delay is around 25 minutes.

The temperature controller, IC2, is driven by sensor IC1, the familiar Type LM35, which produces an output voltage of 10 mV/°C. This voltage is compared with a reference potential provided by a high-stability, temperature-compensated zener diode, D1. Presets P1 and P2 form the fine and coarse controls respectively for setting the tmperature. The comparator switches on T2 whenever the temperature measured by IC1 is below the set value. This causes the relay, Re1, to be energized so that the heater element is powered via the relay contacts.

The timer function is based on oscillator/divider IC3, whose clock frequency is determined by the variable RC network between the PI and PO pins (9 and 11). The clock signal, divided by 2^{10} and 2^{14}, appears at pins 15 and 3 respectively. Toggle switch S1 selects either of these outputs to give time delays of 6 s to 1.5 min. and 1.5–25 min. These settings are marked A and B respectively.

When the delay set by P3-P4 has lapsed, the oscillator in IC3 is disabled by the high level at the pole of the time selector switch. At the same time, T1 is switched on, T2 is switched off, and the (active) buzzer sounds to indicate that the set time has lapsed. The relay is de-energized via T2 and the heater is disconnected from its supply. The timer may be reset while the heater is on by pressing S2

Some accurate calibration is required in the temperature controller. Connect a digital voltmeter between earth and junction R3-P1 and adjust P2 to obtain a voltmeter reading of 100 mV (= 10 °C). Set P1 as appropriate by actually measuring the temperatures at which the relay is energized. Next, set P3 to minimum resistance and S1 to position A and adjust P4 to obtain a time delay of 5–6 seconds after S2 has been pressed. The time delays are set by P3 with the aid of an accurate watch. This procedure is not required for position B, since in this delays 16 times as long as in position A are provided automatically. If a simple thermostat only is required, the timer circuit, T1, T3 and T4 may be omitted.

The circuit is powered by a regulated 5 V supply and draws a current of about 30 mA with the relay inoperative.

The coil resistance of the relay must be not less than 400 Ω.

The temperature sensor must be fitted, of course, at some distance from the heating element.

(C. Sanjay)

052 CAR ELECTRONICS

IMPROVED LOW FUEL INDICATOR

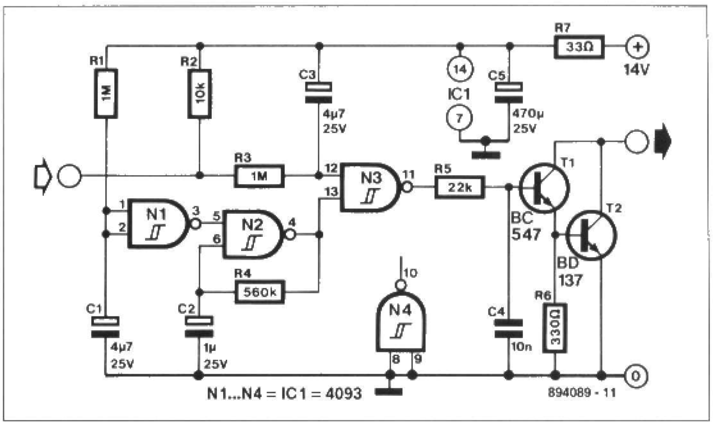
The indicator described here obviates the flickering of the 'low fuel' warning light on the dashboard caused by vehicle movement. The indicator ensures that the light remains extinguished until the duty factor of the signal supplied by the fuel sensor is smaller than 0.5. When that happens, the light comes on and remains on until the fuel tank has been filled to a sufficient level. The present circuit tests the light by causing it to flash for about 5 seconds everyt time the ignition key is turned to start the engine.

The signal processor is switched on together with the ignition. Initially, C1 is discharged, enabling oscillator N2 via inverter N1. One input of N3, pin 12, is connected to R3-C3, the time constant of which is equal to that of R1-C1. If the output of the fuel sensor is high, pin 12 of N3 is held high via R2-C3. The 1.5 Hz signal from oscillator N2 is inverted by N3 and passed to darlington

lamp driver T1-T2. After the delay introduced by R1-C1 has lapsed, gate N1 disables oscillator N2 so that the warning light goes out.

When the fuel sensor inside the fuel tank supplies pulses owing to vehicle movement, C3 is charged via R3 and discharged via R2-R3. When the duty factor (on /off ratio) of the level sensor signal drops below 0.5, the potential across C3 becomes high enough to enable the lamp driver, so that the 'fuel low' warning light is on permanently without any flicker.

(R. Lalic)



053 AUDIO & HI-FI

A SIMPLE VOX

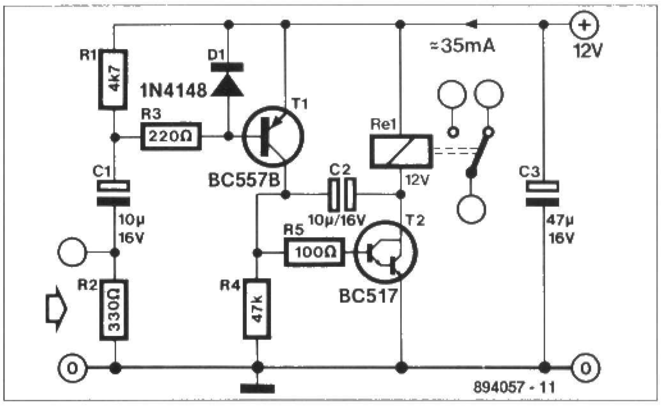
A VOX is a voice-operated switch that is often used as a substitute for the press-to-talk switch on a microphone. The one described here can be connected to almost any audio equipment that has a socket for an external loudspeaker. The actuation threshold is set by the volume control on the AF amplifier that drives the VOX.

The (loudspeaker) signal across R2 is capacitively fed to the base of T1. Resistor R3 limits the base current of this transistor when the input voltage exceeds 600 mV. Diode D1 blocks the positive excursions of the input signal, so that $U_{e\text{-}b}$ can not become more negative than about 0.6 V.

The output relay is driven by darlington T2. Resistor R4 keeps the relay disabled when T1 is off. The value of bipolar capacitor C2 allows it to serve as a ripple filter in conjunction with T2. Resistor R5 limits the base current of T2 to a safe level.

The switching threshold of the VOX is about 600 mV across R2. The maximum input voltage is determined by the maximum permissible dissipation of R2 and R3. As a general rule, the input voltage should not exceed 40 V (p-p).

The current drawn by the VOX is mainly the sum of the currents through the relay coil and through R5. The resistor may carry up to 100 mA when the VOX is overdriven.



(S.G. Dimitriou and F.P. Maggana)

054 AUDIO & HI-FI

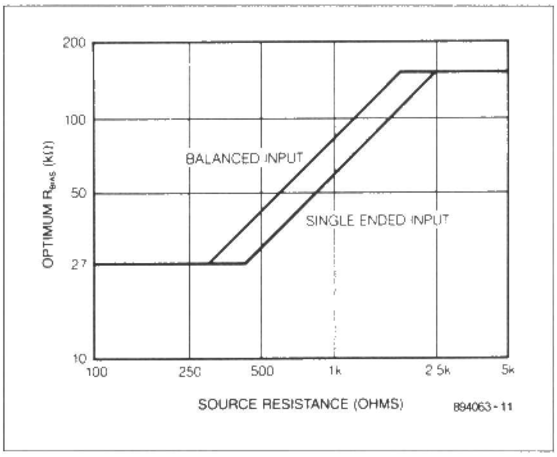
LOW-NOISE MICROPHONE PREAMPLIFIER

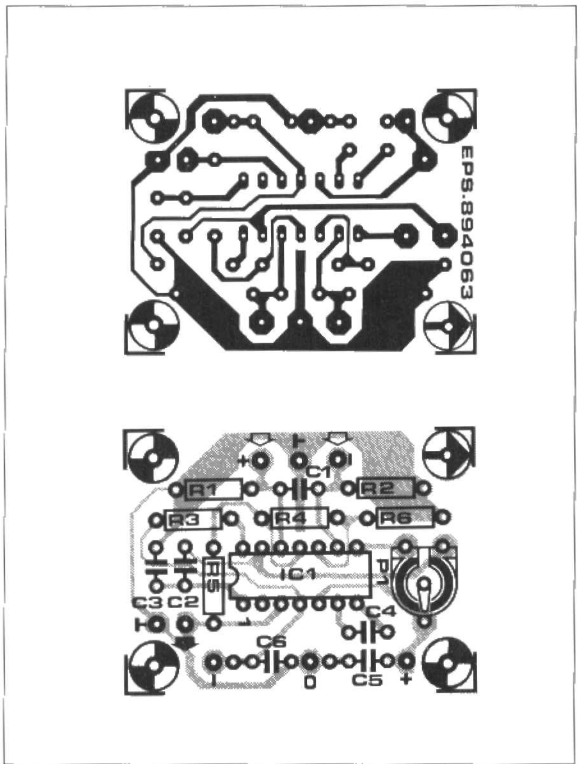
The microphone preamplifier described is based on the SSM2015 from Precision Monolithics Inc. (PMI), which offers a very high gain and very low noise (1.3 nV/√f). It is intended for use with balanced input signals and is capable of providing an amplification of 10–2,000, depending on the value of R4. With R5 = R6 = 10 kΩ, the amplification, A, is calculated from

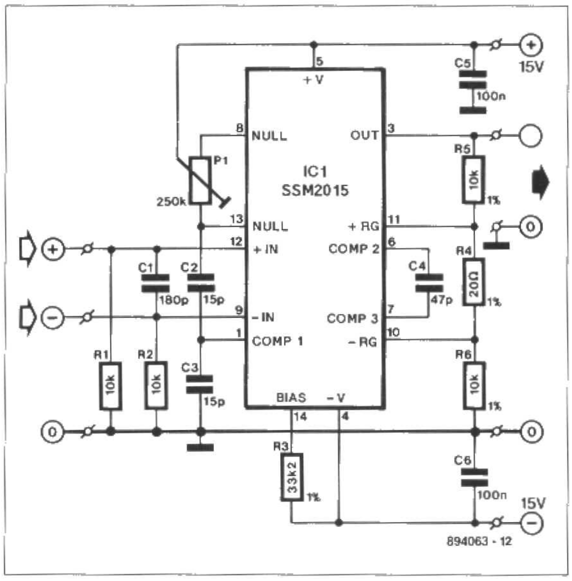
$$A = (20{,}000 / R4) + 3.5$$

With values as shown in the diagram, the amplification is, therefore, about 1,000.

Resistor R3 sets the operating point of the differential input amplifier and thus determines the bandwidth and the slew rate. A value of 33 kΩ gives near-optimum values of these characteristics, but results in a relatively high input bias current of 4.5 µA (with R3 = 150 kΩ , the current is only 1 µA). Moreover, the input noise







Parts list

Resistors:

R1; R2 = 10 k
R3 = 33 k; 1%
R4 = 20 R; 1%
R5; R6 = 10 k; 1%
P1 = 250 k preset

Capacitors:

C1 =180 p styroflex
C2; C3 = 15 p
C4 = 47 p styroflex or ceramic
C5; C6 = 100 n

IC1 = SSM2015

Table 1

R3	C3	C2
27–47 kΩ	15 pF	15 pF
47–68 kΩ	15 pF	10 pF
68–150 kΩ	30 pF	5 pF

Table 2

R3		27–47 kΩ	47–68 kΩ	68–150 kΩ
A = 10	P1 =	500 kΩ	250 kΩ	250 kΩ
A = 100	P1 =	500 kΩ	100 kΩ	100 kΩ
A = 1000	P1 =	250 kΩ	100 kΩ	50 kΩ

level, particularly the current-noise contribution, has increased a little. Nevertheless, the preamplifier has a signal-to-noise ratio of 95 dB, measured with short-circuited + and – inputs and an output level of 0 dBV. Resistor R3 enables the source impedance to be matched to the input of the differential amplifier: if Z = 600 Ω, R3 has an optimum value of 33 kΩ. With a 600 Ω resistor across the input terminals, the signal-to-noise ratio is of the order of 86 dB. The tables and the curve in Fig. 1 give a number of values for the bias resistor and compensation capacitors C2 and C3.

The differential inputs of the SSM2015 are of the floating type, so that external resistors, R1 and R2, are required to give a suitable DC setting. In single-ended (unbalanced) applications, care must be taken to prevent off-sets arising from biasing differences between the inputs, owing to different impedances (ground for one input, and the source for the other). Resistors R1 and R2 cause common-mode noise and must not be given a higher value than shown in the diagram.

Off-set compensation with the aid of preset P1 is required for the given value of R3, which results in the available amplification of 1,000. The value of R3 depends on the gain setting – see Table 2. Capacitor C4 compensates the on-chip input current regulator, and C1 suppresses HF signals.

Distortion of the preamplifier at 1 kHz and 0 dBV was measured at less than 0.006% and less than 0.01% at a test frequency of 10 kHz. The half-power bandwidth is 180 kHz at 3 V across 1 kΩ. Common-mode rejection at 50 Hz is greater than 100 dB.

PMI states that the output of the SSM2015 is not intended to drive long lines: capacitive loads greater than 150 pF should be decoupled by a 100 Ω resistor in series with the output (but note that R5 must remain connected to pin 3).

(Precision Monolithics Inc.)

055 GENERAL INTEREST

HIGH-VOLUME ALARM

When this alarm is actuated by a low-level signal at input $\overline{\text{EN}}$, the (HF) loudspeaker produces a number of 4-tone sequences separated by quiet intervals. Each sequence sounds louder than the previous one to give the alarm a very distinctive character. The peak output is reached after about 28 seconds.

As long as the $\overline{\text{EN}}$ input is logic high, counters IC1a and IC1b remain reset and interval oscillator N2 and tone generator N3 are disabled. The alarm is then off.

When $\overline{\text{EN}}$ is actuated, the oscillator and the two counters are enabled. Counter IC1a is clocked with pulses (PRF = 8 Hz) from N2. Gates N1 and N4 at counter outputs Q0 and Q3 cause T1 to be turned off during eight consecutive clock cycles from IC1a.

During the next eight cycles, the transistor is alternately switched on and off as illustrated in the timing diagram. The loudspeaker sounds only when T1 conducts.

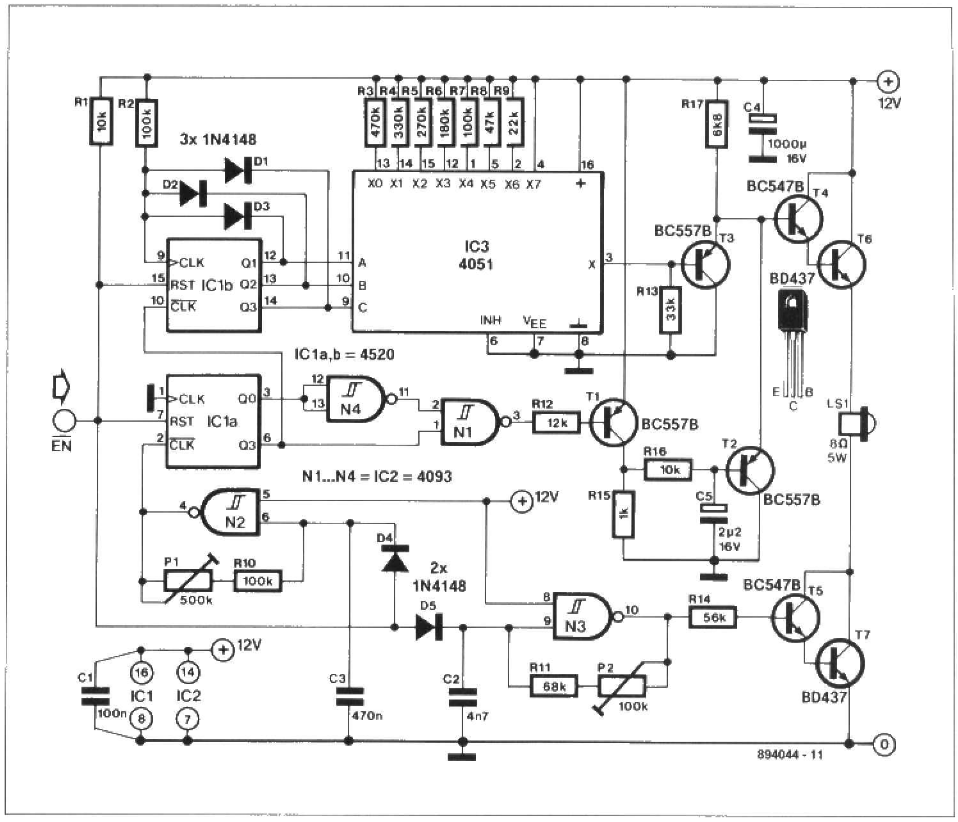
Since output $\overline{Q3}$ of counter IC1a is connected to the CLK input of counter IC1b, the latter is incremented by the 16th negative pulse transition. In practice, this means that counter IC1b is clocked after each tone sequence. The most significant outputs of counter IC1b drive the 3-bit selection inputs of analogue multiplexer IC3. Since Q0 is not used, IC1b requires two clock pulses to enable the multiplexer to connect the next input, Xn, to the output, X. The seven resistors at the multiplexer inputs cause the base voltage of T3 to increase after each alternate tone sequence. As a result, the voltage across the loudspeaker increases, so that the alarm sounds louder.

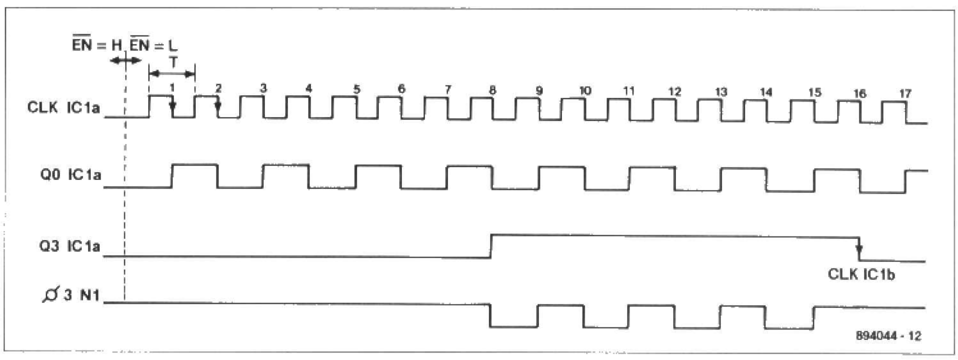
Transistor T1 switches the successive bleeps on and off. Resistor R16, capacitor C5 and transistor T2 prevent abrupt switching of the voltage across the loudspeaker when T1 is turned off, and ensure that the sound level reverts slowly to the level set by the resistor at the relevant multiplexer input.

After the alarm is actuated, the output volume rises sevenfold. Diodes D1, D2 and D3 cause counter IC1b to stop at state 1110, so that the multiplexer passes the full positive supply voltage at input X7 to volume control T3. The alarm then sounds continuously at maximum volume, requiring the power supply to provide a peak current of up to 1.25 A through the loudspeaker. The resultant sound is ear-piercing.

Prtesets P1 and P2 serve to set the interval repetition rate and the sound frequency respectively.

(C. Sanjay)





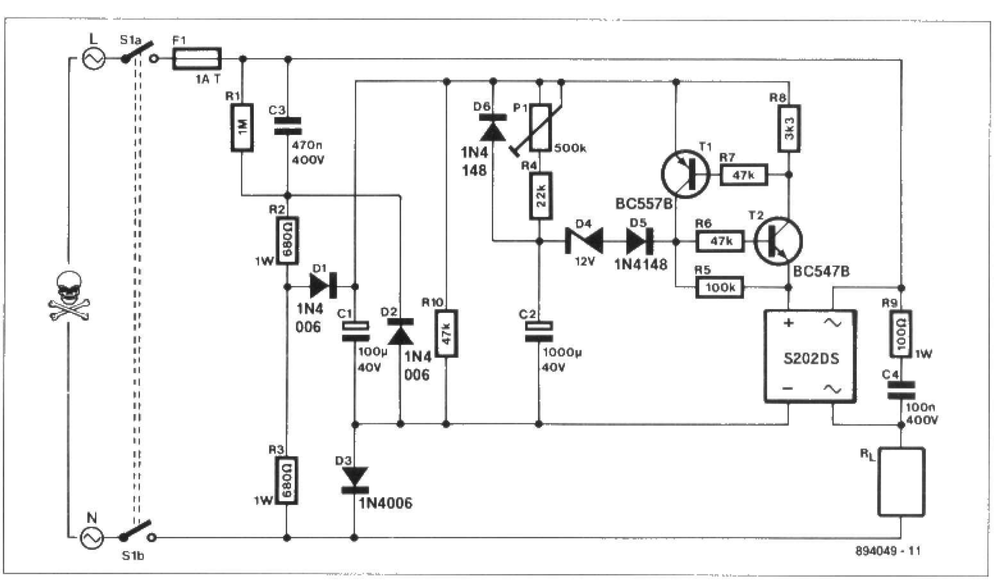
056 GENERAL INTEREST

MAINS POWERED TIMER

This timer may be inserted in a power line to provide a controllable delay before a load is energized. It was developed to work in conjunction with a passive infrared movement detector as part of an intruder alarm.

The mains voltage is reduced by C3 and rectified to give about 30 V across C1. This potential charges C2 slowly via R4-P1. When U_{C2} reaches about 14 V, electronic switch T1-T2 actuates a solid-state relay (a Type S202DS from Sharp). When the mains voltage is removed,C2 discharges rapidly via D6 and R10. The delay extends from 15 s (P1 set to minimum resistance) to 5 min (P1 set to maximum resistance).

The solid-state relay needs cooling in accordance with the current drawn by the load: at up to 1 A no heat sink is re-



quired; at 1–3 A (max), a 5×5 cm heat sink is advisable.

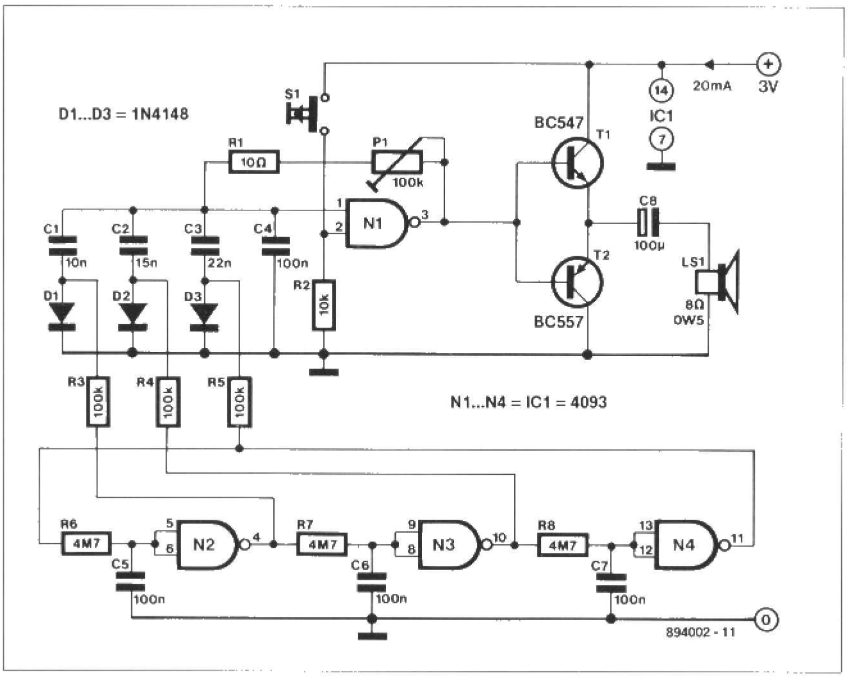
During the building of the circuit, due consideration must be given to safety since many parts will be at mains potential. For instance, fitting the unit in an ABS or other man-made fibre enclosure is a must. If a potentiometer is used for P1, its spindle should be an insulated type. If a preset is used, it must not be accessible through a hole in the enclosure.

Switch S1 is a DPST type that disconnects the circuit from the mains. Nevertheless, the only way of working on the circuit in safety is by taking the plug out of the mains socket and allowing C3 sufficient time to discharge.

(D. Dwyer)

057 GENERAL INTEREST

SINGLE-CHIP MELODY GENERATOR



This melody generator, based on a Type 4093 CMOS Schmitt trigger, may be used in alarms, doorbells and cars (audible reverse gear or lights on indicator).

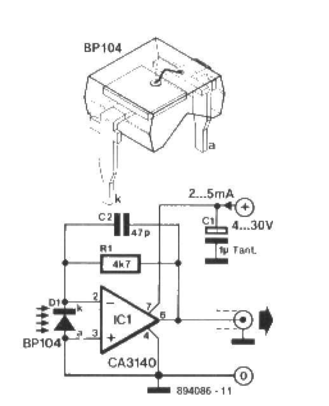
Three of the four NAND gates in the 4093 are connected in series by RC networks. Oscillation is effected by feedback of the output signal of N4 to the input of N2. The logic high levels produced by the cascaded gates in the oscillator circuit are used for biasing one of associated diodes D1, D2 and D3. The relevant diode connects one of frequency-determining capacitors C1–C3 to tone oscillator N1. The audio signal available when S1 is pressed is applied to complementary transistor pair T1-T2 that drives the loudspeaker.

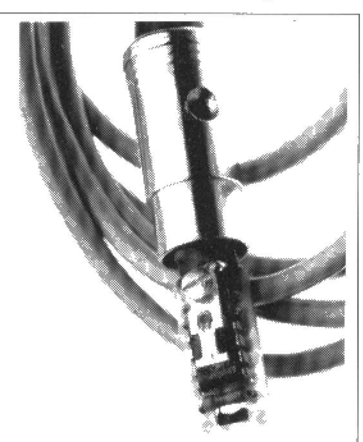
The frequency of the emitted tone may be adjusted to individual taste by preset P1.

058 GENERAL INTEREST

INFRA-RED MICROPHONE

The circuit of this microphone was originally designed to monitor a 7-segment display used on the flight deck of a Boeing 737. That display used filaments, so that an IR detector was an obvious choice. In the Boeing, it was connected to a small portable recorder: the sensor therefore acted as a microphone that reacted to IR light instead of to sound. This idea of an IR microphone is made more tangible by housing the device in the shell of DIN plug as shown in the photograph.





The IR microphone consists of a Type BP104 photodiode connected to the inputs of a DC-coupled operational amplifier, whose gain is determined by R1.

The device may be used to 'listen' to the visual world around us. It is particularly effective where sources of noise, such as incandescent light bulbs, are switched off. A gas flame, such as that of a cigarette lighter, is manifested as a soft breeze. A cosy fire burning in the grate comes out as a real hurricane. This means that the microphone may be used as an acoustic fire alarm, but that is about the only application we can think of. However, the circuit is intended more to give us an opportunity to see our environment from a different angle. If the BP104 is replaced by a BPW34, the sensitivity of the device is shifted from the infra-red to the visible spectrum.

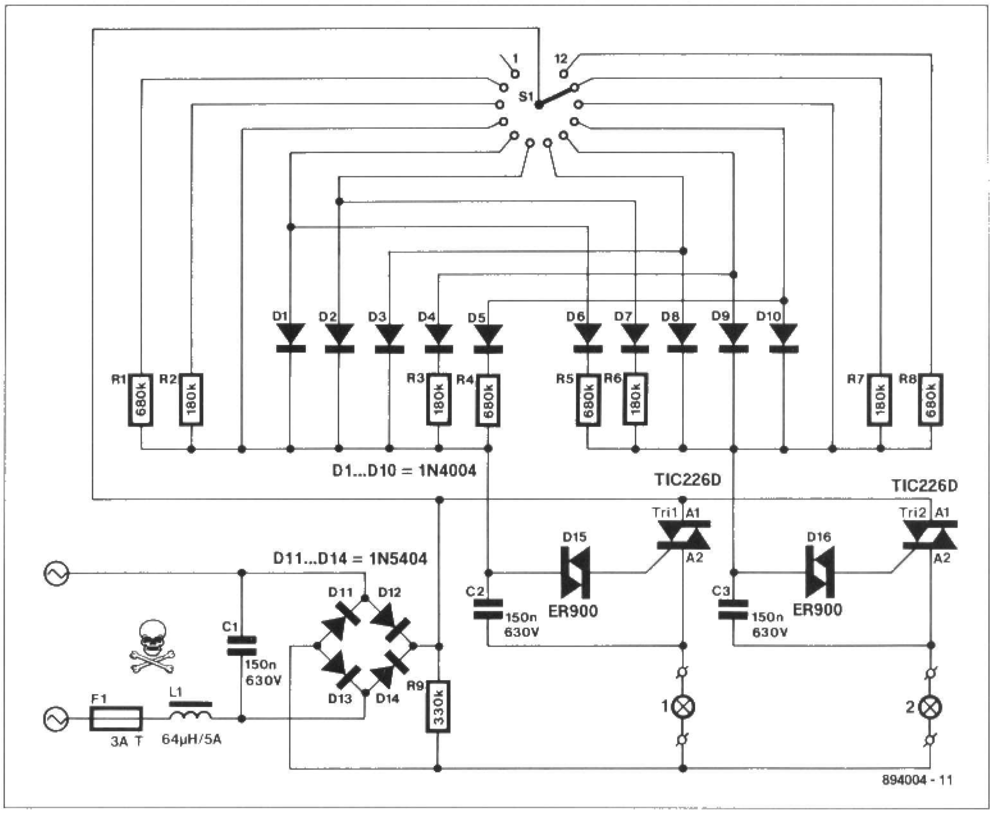
The current drawn by the circuit depends to some extent on the supply voltage and should be 2–5 mA.

059 GENERAL INTEREST

FOUR-QUADRANT DIMMER

This very special mains-operated dimmer for domestic or industrial lights is not available in proprietary form: it enables brightness control of two groups of lights in one operation. The possible combinations of brightness are shown in the table. It will be clear that it is not possible to obtain continuous control of brightness in the two groups. Instead, the circuit affords the setting of four states of brightness in either group: full on, fully dim-med, 1/3 on and 2/3 on.

Both sections of the circuit operate on the well-known principle of a triac being switched from the blocking state to the conducting state with the aid of an RC network and a diode. The RC network provides the necessary phase shift and determines when the triac is switched. The rotary switch selects the resistor in a given network and thus the brightness of the relevant group of lights. No resistor means that the group is off; a short-circuit gives maximum brightness, and resistors of 10 kΩ and 18 kΩ mean intermediate brightness. The diodes prevent the groups affecting one another.



The 64 µH choke, L1, and the 150 nF capacitor across the bridge rectifier prevent the dimmer causing interference in other equipment connected to the mains.

If the triacs are fitted on a heat sink rated at 12 K/W, up to 500 W per group may be controlled. It is, of course, essential that the enclosure in which the dimmer is fitted provides ample cooling: a fair number of slots or holes in it are, therefore, essential: these should not permit the circuit elements to be touched.

The switch should have a non-metallic spindle: this is not only safer than a metallic one, but it also enables the easy removal of the end-notch so that the switch may be rotated continuously instead of having to be returned to the first stop every time it is operated.

It is recommended that mains on-off switch S2 is fitted with a built-in 'on' indicator bulb: this shows at a glance whether the circuit is on even though S1 may be in the off position .

Finally, do bear in mind that this circuit carries mains voltage in many places: good workmanship and insulation are, therefore, of the utmost importance.

(C.G. Mangold)

Switch position	Brightness	
	Group A	Group B
1	0	0
2	1/3	0
3	2/3	0
4	1	0
5	1	1/3
6	1	2/3
7	1	1
8	2/3	1
9	1/3	1
10	0	1
11	0	2/3
12	0	1/3

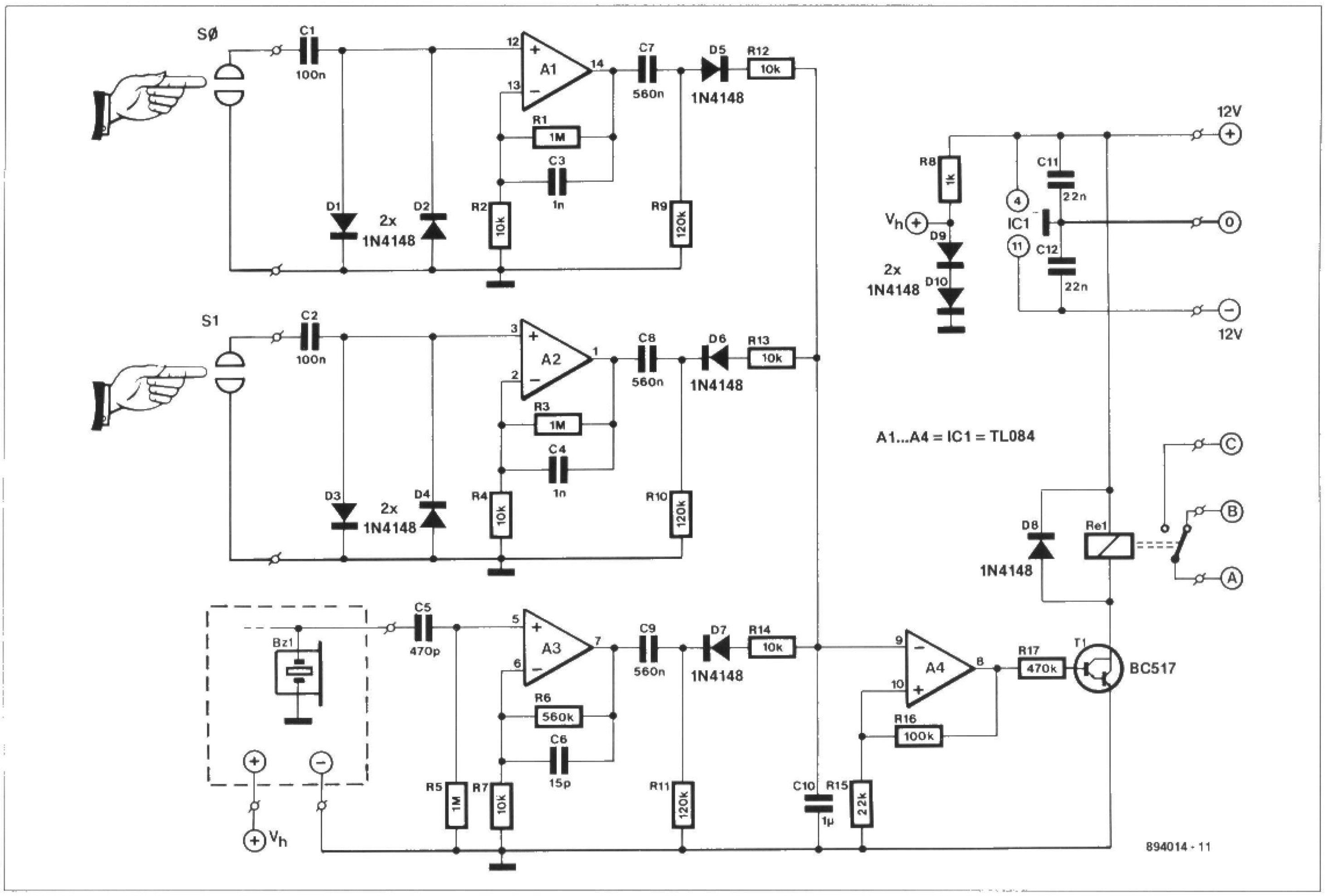
060 GENERAL INTEREST

SENSOR SWITCH AND CLOCK

One Type TL084 IC and an old quartz watch enable the construction of a de luxe on-off switch. Two of the four opamps contained in the TL084 (A1 and A2) are used to amplify the input signals from the sensors hundredfold (with the component values as shown in the diagram). Just touching the sensors with a finger causes a good 50 Hz input signal (hum). Note that the amplification drops rapidly with rising frequency.

Diodes D5 and D6 rectify (single-phase) the 50 Hz signal. Since the diodes are connected in anti-phase, touching the 'off' sensor results in a positive potential across C10, whereas touching the 'on' sensor gives a negative potential across C10.

Opamp A4 is connected as an inverting bistable, so that a nega



tive potential across C10 causes relay Re1 to be energized. Because of feedback resistor R16, this state is maintained until the other sensor is touched.

The relay may also be energized at a predetermined time with the aid of a quartz watch. The 1.2 V supply for the watch is derived from the voltage drop across diodes D9 and D10: it may be increased to 1.8 V by adding a third diode.

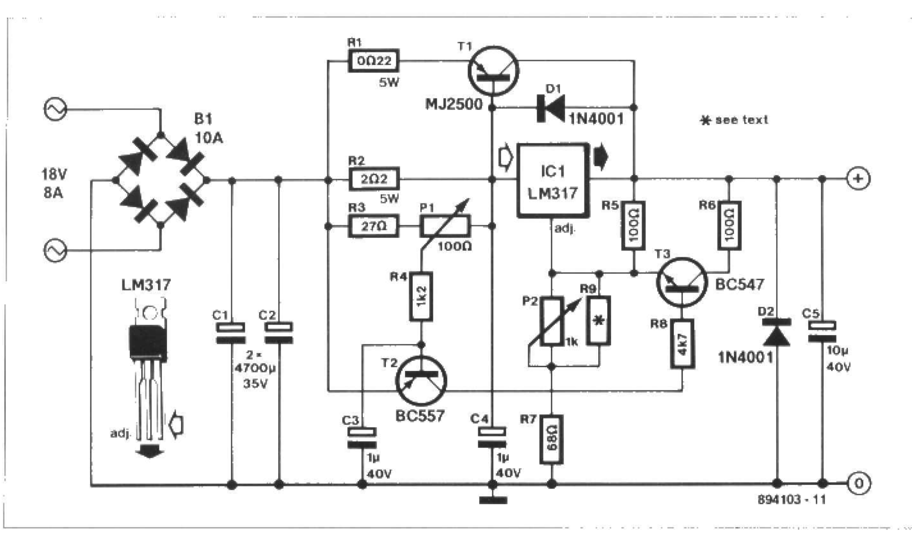
The piezo buzzer in the watch is connected to the input of A3 via C5. As soon as the alarm goes off (the hour signal must be off), the voltage across C10 becomes negative, the relay is energized and the load is switched on.

The circuit, excluding the relay, draws a current of about 20 mA.

(R. Ochs)

061 GENERAL INTEREST

MINI-DRILL CONTROL



The circuit described here is intended as a revolution control for small DC motors as fitted, for instance, in small electric drills (such as used in precision engineering and for drilling printed-circuit boards, among others). The behaviour of these motors, which are normally permanent magnet types, is comparable to that of independently powered motors.In theory, the RPM of these motors depends solely on the applied voltage. The motor adjusts its RPM until the counter EMF generated in its coils is equal to the applied voltage. There is, unfortunately, a drop across the internal resistance of the motor and this causes the RPM to drop in relation to the load. In other words, the larger the load, the larger the drop across the internal resistance and the lower the RPM.

The present circuit provides a kind of com-

pensation for the internal resistance of the motor: when the current drawn by the motor rises, the supply voltage is increased automatically to counter the fall in RPM.

The circuit is based on an enhanced voltage regulator consisting of IC1 and T1, which provides a reasonably large output current (even small drills draw 2–5 A). The 'onset' supply voltage, and thus the RPM, is set by P2. Because of emitter resistor R1, the currents through IC1 and T1 will be related to one another in the ratio determined by R1 and R2. Owing to this arrangement, the internal short-circuit protection of IC1 will also, indirectly, provide some protection to T1.

As soon as the current drawn exceeds a certain value, T2 will be switched on. This results in a base current for T3 so that R5 is in parallel (well, more or less) with R6. This arrangement automatically raises the output voltage to counter a threatened drop in RPM. The moment at which this action occurs is set by P1, so that the present circuit can be adapted pretty precisely to the motor used.

If only very small motors are likely to be used, the power supply (transformer and bridge rectifier) may be rated rather more conservatively. As a guide, the current in the transformer secondary should be about one and a half times the maximum DC output current.

(G.J. Lammertink)

062 RADIO & TV

CALL TONE GENERATOR

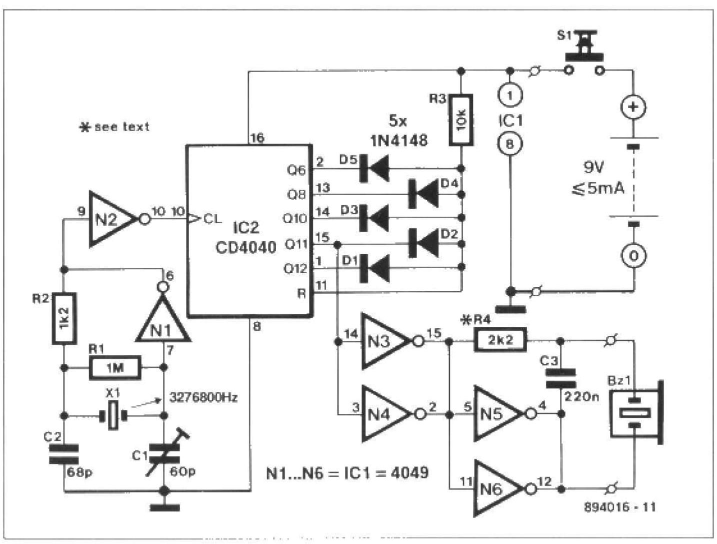
Amateur VHF relay stations are normally actuated by a 1750 Hz call tone. This may give problems when the relevant sending equipment has no internal call tone generator, or it has one whose frequency is not sufficiently accurate, or whose tone duration is not long enough to securely energize the relevant relay.

These problems can be overcome by the stand-alone generator described here. Simply placed in front of the microphone, it makes absolutely certain that the relay station is actuated.

The generator consists of a quartz oscillator, a frequency counter and a buffer-amplifier, all contained in just two CMOS ICs. It is powered by a 9 V PP3 battery, from which it draws a current of around 5 mA.

Gates N1 and N2 form an oscillator that is controlled by a 3.27680 MHz crystal and provides clock pulses to IC2 which is connected as a programmable scaler. Diodes D1–D5 determine the divide factor of 1872. Counter output Q1 thus provides the wanted 1750 Hz signal, which is buffered by N3–N6 before being applied to a piezo electric buzzer. Capacitor C3 suppresses any harmonics, while R4 determines the volume of the output signal.

(N. Körber)



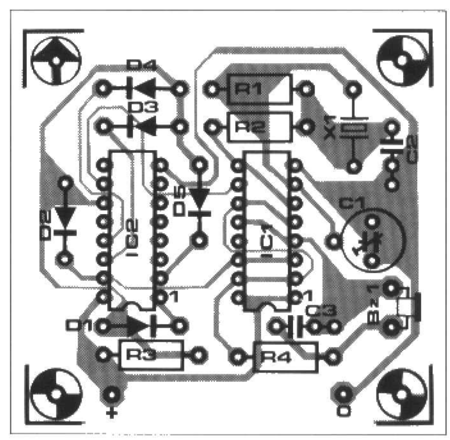
Parts list

Resistors:
R1 = 1 M
R2 = 1k2
R3 = 10 k
R4 = 2k2

Capacitors:
C1 = 60 pF trimmer
C2 = 68 p
C3 = 220 n

Semiconductors:
D1–D5 = 1N4148
IC1 = 4049
IC2 = 4040

Miscellaneous:
X1 = 3.2768 MHz; 30 p parallel
Bz1 = piezo electric buzzer

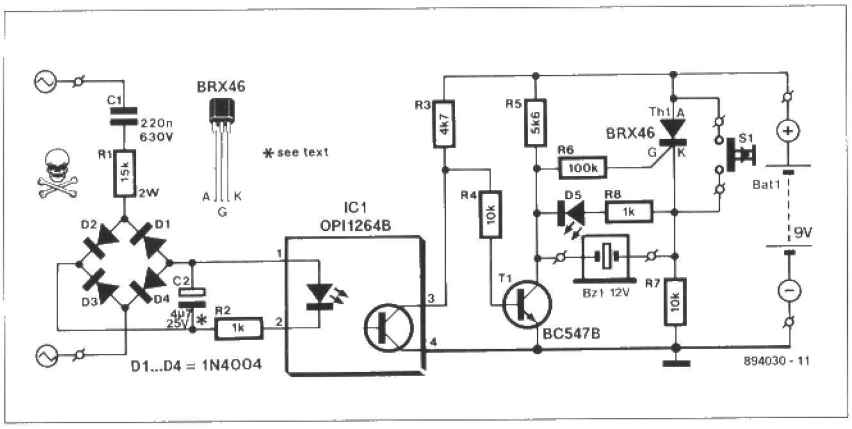


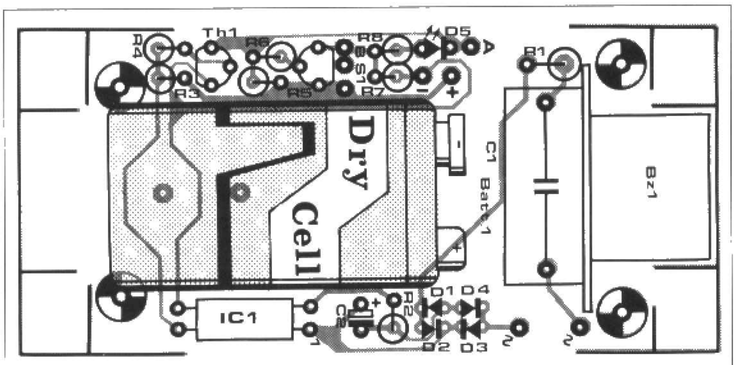
063 GENERAL INTEREST

MAINS FAILURE INDICATOR

When the mains voltage is present at the input terminals, the transistor in the optocoupler is on, T1 is off and silicon-controlled rectifier Th1 is in the conducting state. Since both terminals of the piezo electric buzzer are then at the same potential, the buzzer is inactive. If the mains voltage drops out, transistor T1 conducts and causes one of the terminals of the buzzer to be connected to earth; the thyristor remains in the conducting state. In this situation there is a large enough potential difference across both the buzzer and D5 to cause these elements indicating the mains failure both audibly and visibly.

When the mains is restored, the circuit returns to its original state. A touch on the reset button then interrupts the current through the SCR so that the thyristor goes into the blocking state, and the other terminal of the buzzer is connected to ground.





The unit is powered by a 9 V PP3 battery and draws a quiescent current of 1.7–2.5 mA.

It is important that the enclosure is a well-insulated type.

Finally two points to note. If by accident the circuit to the optocoupler and R2 is broken, electrolytic capacitor C2 may be damaged since it will be charged well above its 25 V rating. Secondly, where a plug is used for the mains connection, it is advisable to solder a 1MΩ resistor across C1 so that this capacitor does not retain its charge after the plug is removed from the mains socket.

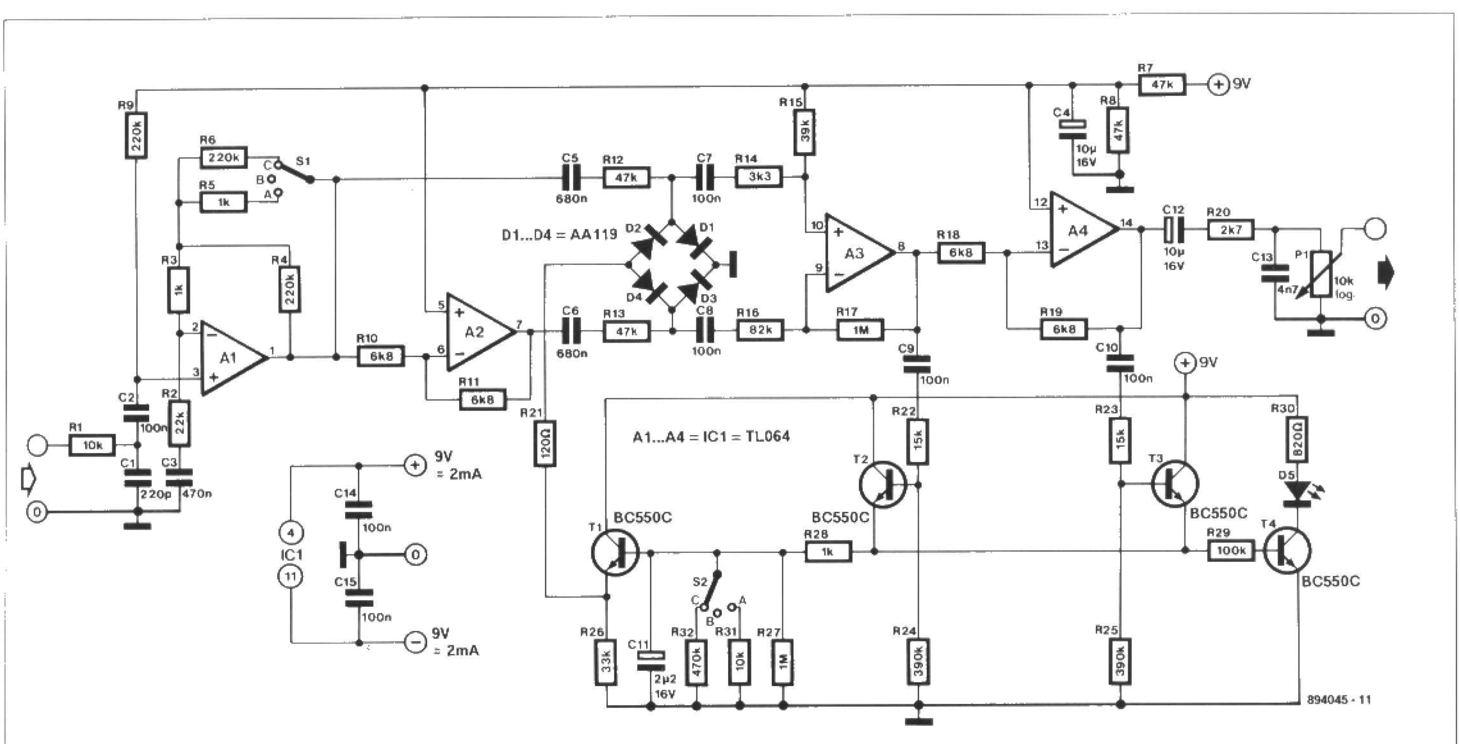
Parts list	
Resistors:	Semiconductors:
R1 = 15 k; 2 W	D1–D4 = 1N4004
R2; R8 = 1 k	D5 = LED
R3 = 4k7	T1 = BC547B
R4; R7 = 10 k	Th1 = BRX46
R5 = 5k6	IC1 = OPI 1264B
R6 = 100 k	
	Miscellaneous:
Capacitors:	S1 = switch with 1 make contact
C1 = 220 n; 630 V	Bz1 = piezo buzzer 9 V
C2 = 4µ7; 25 V	9-V PP3 battery

064 ELECTROPHONICS

GUITAR COMPRESSOR

The control of this compressor is based on the dependence of the dynamic resistance of a diode on the current flowing through it. The heart of the present circuit is the diode bridge D1–D4, which behaves as a variable resistance controlled by the current flowing in T1.

The input signal is applied to preamplifier stage A1 via low-pass filter R1-C1 that removes any HF noise from the input. Switch S1 in the feedback loop of A1 sets the amplification to 1 (position



A), 6 (C) or 11 (B). The amplified signal is applied to the diode bridge direct via R12 and C5, and inverted via inverter A2, capacitor C6 and resistor R13. The two signals are summed by the bridge, amplified (in A3) and then split again into two, one of which is inverted by A4. The positive half cycles of the two signals are used to switch on T2 and T3 respectively. Capacitor C11 is then charged via R12. When the potential across this capacitor reaches a certain level, T1 is also switched on, after which a control current flows through the bridge via R21. This current lowers the resistance of the bridge so that the signal is attenuated (compressed). At the same time, the LED lights to indicate that the signal is being compressed. Capacitor C12 prevents any DC voltage from reaching the output.

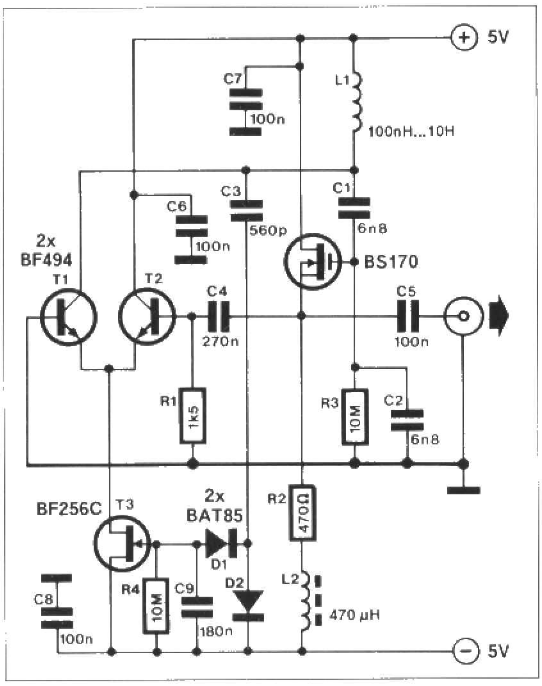
The output signal is taken from the wiper of P1. Low-pass section R20-C13 limits its bandwidth to 12 kHz.

Switch S2 enables the selection of various decay times of C11. The values shown in the diagram have in practice proved to be the most useful. Nevertheless, these values are subjective and may be altered to personal taste and requirements.

(W. Teder)

065 TEST & MEASUREMENT

LC SINE WAVE GENERATOR



This compact LC oscillator offers a frequency range of about 1 kHz to almost 9 MHz and a low-distortion sine wave output.

The heart of the circuit is series-resonant circuit L1-C2-C3 in the feedback loop of amplifiers T1-T2. Transistor T2, which is connected as an emitter follower, serves as impedance converter, whereas T1, connected in a common base circuit, is a voltage amplifier whose amplification is determined by the impedance of L1 in its collector circuit and the emitter current. The feedback loop runs from the collector of T1 via the junction of capacitive divider C1-C2, source follower BS170 and the input impedance formed by R1 and C4. The whole is strongly reminiscent of a Colpitts circuit. The signal is also taken to the output terminal via C5.

Of particular interest is the amplitude control by the current source. The signal is rectified by two Schottky diodes, smoothed by C9 and then used to control the current through T3. The gain of amplifier T1 is therefore higher at low input levels than at higher ones. This arrangement ensures very low distortion, since the amplifier can not be overdriven.

The resonant frequency may be calculated from

$$f = 1 / 2\pi \sqrt{[L1\ C1\ C2 / (C1 + C2)]}$$

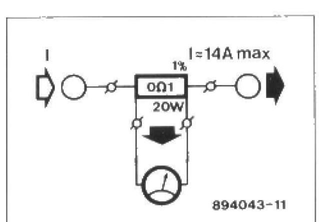
With values as shown, it extends from 863 Hz ($L1$ = 10 nH) to 8.630 MHz ($L1$ = 100 nH).

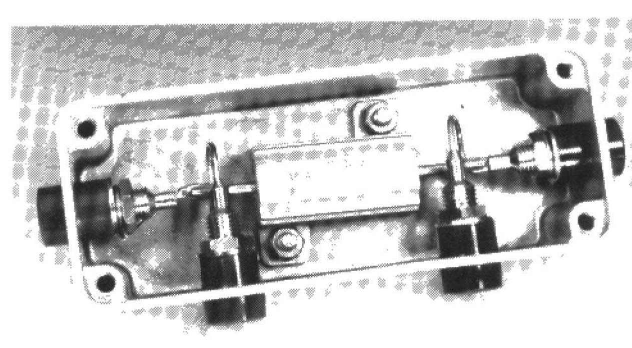
The unit may be used to measure the Q of inductors. To that end, a potentiometer is connected in parallel with L1 and adjusted so that the current through the amplifier is doubled. The Q is then calculated from

$$Q = Rp / 2\pi f L$$

066 TEST & MEASUREMENT

SHUNT FOR MULTIMETER





The current range in multimeters, particularly the more inexpensive ones, is restricted by the load limits of the internal shunts to 1–2 A.The photo shows how easily a precision heavy-duty resistor from Dale or RCL (0.1 Ω; 20 W; 1%) may be used as an external shunt. These resistors were not designed for this purpose, but they are much cheaper than custom made shunt resistors. The 20 W rating applies only, by the way, if a heat sink is used: without that its rating is only 8 W.

The maximum current through the device on a heat sink is about 14 A; the larger versions draw up to17.5 A. When mounting the shunt, make sure that the test terminals as well as the device terminals are soldered properly, otherwise the resistance of the terminals is added to the shunt.

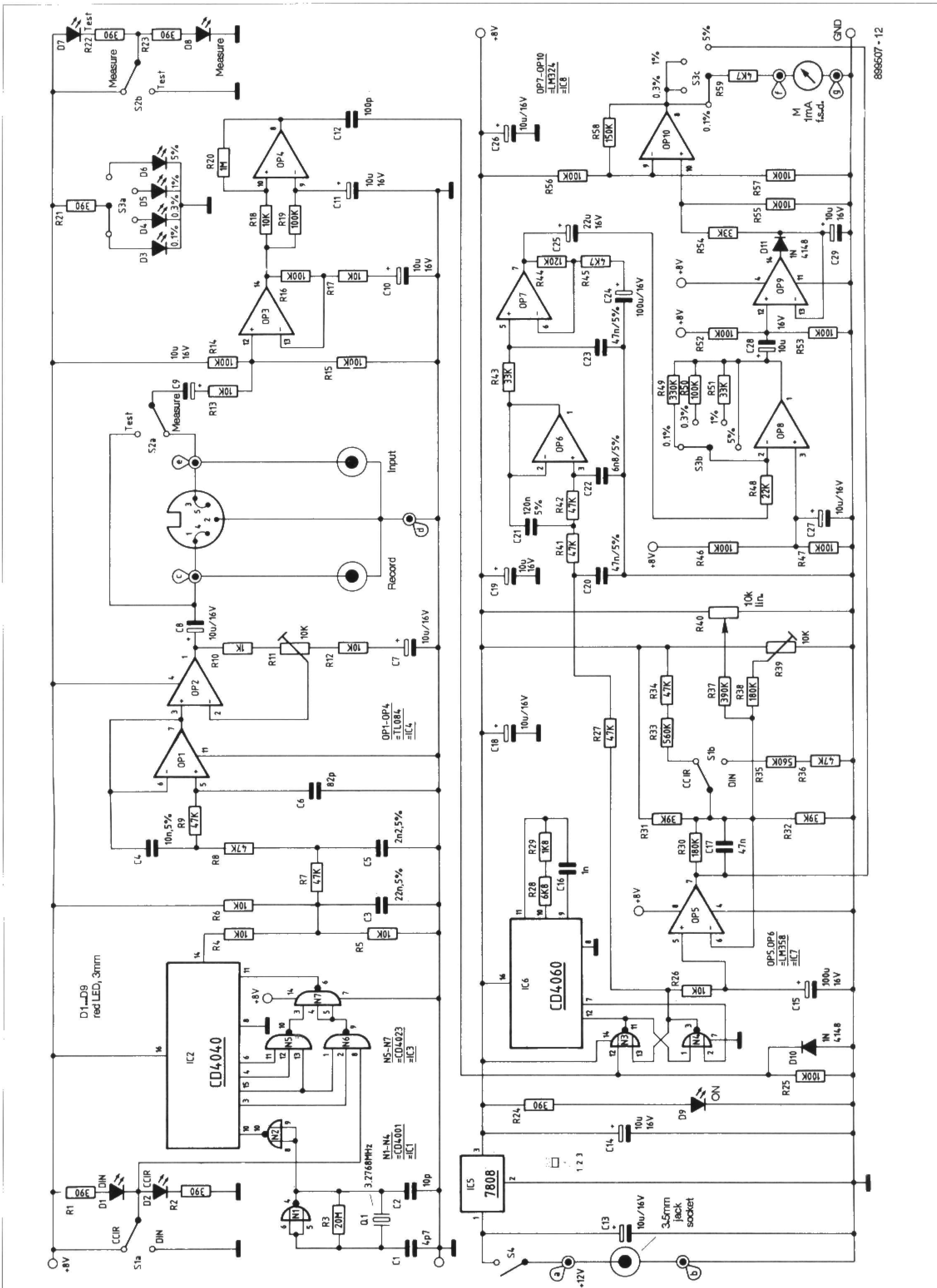


Fig. 2. Circuit diagram of the tracking tester.

tified signal is shifted from half the supply potential to ground by OP_{10}, so that tracking measurements can use the full meter scale starting at 0%.

The centre-zero indication for absolute tape speed measurements is obtained with the aid of integrating network R_{26}-C_{15} and an amplifier based on OP_5. In the TEST mode, the meter is set to the centre indication with R_{39} (coarse) and R_{40} (fine). The measuring range is ±5%.

Setting up

The tracking tester can be adjusted with a frequency meter and an oscilloscope.

To begin with, preset R_{11} is set for an output level of 775 mV_{rms} at pin 1 of OP_2. Other reference levels may be set where these are required.

The CALIBRATE control is set to the centre of its travel. Switch S_2 is set to TEST, and preset R_{39} is adjusted until the meter indicates 0% at the centre of the scale. Range switch S_3 is set to 'Drift 5%' for this adjustment.

When the centre indication can not be achieved by turning preset R_{39}, the deviation of the oscillator frequency generated by IC_6 is too large. This frequency should be 49.2 kHz nominally with a maximum tolerance of ±10%. It can be measured at pin 9 of IC_6, while pin 12 of gate N_3 is provisionally made logic high with the aid of a short wire to the +8 V supply rail. Make small changes to R_{29} and/or R_{28} to pull the frequency within the acceptable range. Two series-connected resistors are used here for this setting instead of the more usual trimmer capacitor, which would magnify the effect of stray capacitance. The minimum equivalent resistance of R_{28}-R_{29} must not be made lower than 6.8 kΩ, and not higher than 22 kΩ. The wire connection at pin 13 of N_3 is removed after the necessary corrections have been made.

When a frequency meter is not available, R_{28} and R_{29} may be changed in 1 kΩ increments until the centre-zero indication is achieved within the range of R_{39}.

The above corrections are not likely to be necessary in most cases, since the circuit is dimensioned such that R_{39} covers the required control range. The frequency correction with the oscillator components around IC_6 is described for the sake of completeness.

No other adjustments are required.

Construction

The circuit is built on two printed circuit boards, the main board and the oscillator board. The component overlays in Fig. 3 and the Parts List are given as an aid for populating the boards, which should be started with the low-profile components. All soldering is done at the track sides.

The main board has some components that are mounted at the track side: voltage regulator IC_5, electrolytic capacitors C_{15}, C_{24}, C_{25} and all solder terminals.

After a thorough check of the completed boards, the oscillator board may be mounted upright on to the main board in a manner where the copper surfaces at the side of D_{10} line up with the associated surfaces on the main board. The copper surfaces are then joined with plenty of solder to secure the oscillator board at right angles on to the track side of the main board. The oscillator board is relatively small and light, so that additional mechanical support is not required.

The two phono sockets and the DIN socket are secured on to the front panel, which also serves to hold the main board by means of the nuts on the threaded shafts of the switches. One nut is first turned on to each of the shafts, followed by a locking washer, after which the spindles are inserted into the holes provided in the front panel. Finally, the PCB assembly is secured to the front panel by one additional nut for each protruding switch shaft.

The spindles of the potentiometer and the range switch are then cut to length and fitted with the associated knobs.

Finally, the wiring is installed in accordance with the circuit diagram. This includes the connection of the moving-coil meter and the supply voltage. The latter is applied to the circuit via a 3.5 mm jack socket on the rear panel of the enclosure. The recommended input voltage is 9 VDC.

The meter is secured to the inside of the front panel. This requires removing a part of the rims at the inside of the top and bottom halves of the enclosure to ensure that the face of the meter is flush with the front panel. The meter is carefully secured with a some two-component adhesive or super-glue applied at the corners.

A complete kit of parts for the Tracking Tester, which is designed in West-Germany, is available from the designers' exclusive worldwide distributors (regrettably not in the USA and Canada):

ELV France
B.P. 40
F-57480 Sierck-les-Bains
FRANCE
Telephone: +33 82827213
Fax: +33 82838180

Parts list

Resistors:
R_1;R_2;R_{21}–R_{24}=390R
R_{10}=1K0
R_{29}=1K8
R_{45};R_{59}=4K7
R_{28}=6K8
R_4;R_5;R_6;R_{12};R_{13};R_{17};R_{18};R_{26}=10K
R_{48}=22K
R_{43};R_{51};R_{54}=33K
R_{31};R_{32}=39K
R_7;R_8;R_9;R_{27};R_{34};R_{36};R_{41};R_{42}=47K
R_{14};R_{15};R_{16};R_{19};R_{25};R_{46};R_{47};R_{50};R_{52};
R_{53};R_{55};R_{56};R_{57}=100K
R_{44}=120K
R_{58}=150K
R_{30};R_{38}=180K
R_{49}=330K
R_{37}=390K
R_{33};R_{35}=560K
R_{20}=1M0
R_3=20M
R_{11};R_{39}=10K preset H
R_{40}=10K linear potentiometer; spindle dia. 6 mm

Capacitors:
C_1=4p7
C_2=10p
C_6=82p
C_{12}=100p
C_{16}=1n0
C_5=2n2; 5%
C_{22}=6n8; 5%
C_4=10n; 5%
C_3=22n; 5%
C_{17};C_{20};C_{23}=47n; 5%
C_{21}=120n; 5%
C_7–C_{11};C_{13};C_{14};C_{18};C_{19};C_{26}–C_{29}=10µ; 16 V
C_{25}=22µ; 16 V
C_{15};C_{24}=100µ; 16 V

Semiconductors:
IC_4=TL084
IC_8=LM324
IC_7=LM358
IC_1=CD4001
IC_3=CD4023
IC_2=CD4040
IC_6=CD4060
IC_5=7808
D_{10};D_{11}=1N4148
D_1–D_9= LED; 3 mm; red

Miscellaneous:
Q_1 = 4 MHz quartz crystal.
S_3 = 3-pole, 4-way rotary switch for PCB mounting.
S_1;S_2 = miniature DPDT switch.
S_4 = miniature SPDT switch.
10 solder terminals.
30 cm screened wire, single core.
20 cm screened wire, 2-core 0.4 mm^2.

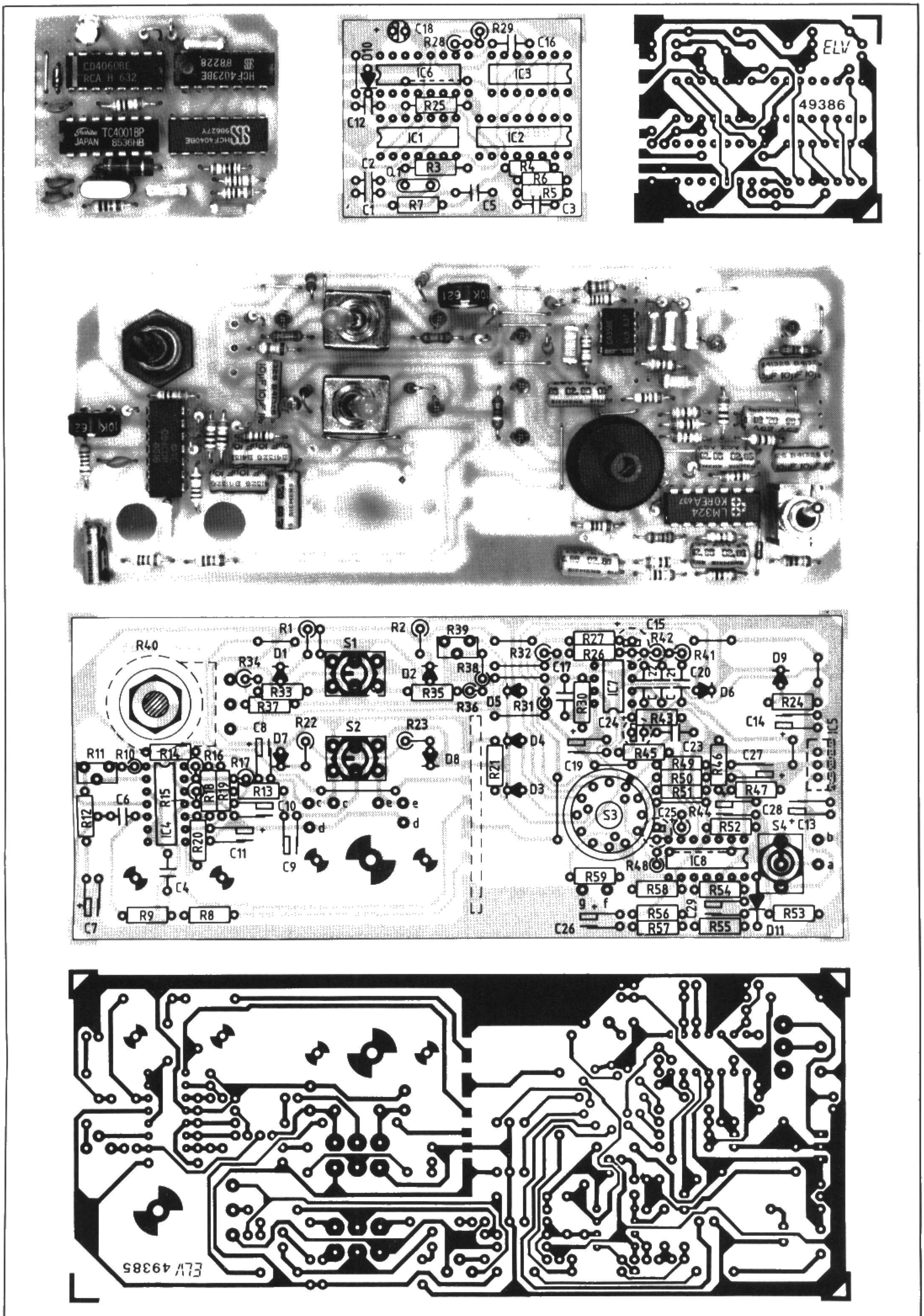


Fig. 3. Top views and component overlays of the oscillator board and the main board of the instrument.

APPLICATION NOTES

The contents of this article are based on information obtained from manufacturers in the electronics industry, or their representatives, and do not imply practical experience by *Elektor Electronics* or its consultants.

VIDEO COMPRESSION/EXPANSION PROCESSOR

The recently introduced AM95C71 from Advanced Micro Devices compresses and expands bit-mapped images for storage and processing in facsimile equipment and, of course, personal computers. The device is a significant step forwards towards the paperless office, and PC users will immediately see opportunities to design a really fast hardware frame grabber.

Scanned images contain much white space and redundant information. If the image is coded to eliminate redundancy, it requires less memory when stored, and can be transmitted at a given data rate in significantly less time. Efficiency is important in today's information-hungry applications, even in high-bandwidth environments. With low data rates, efficacy becomes critical and inefficiency leads to bottlenecks. An example of this is when an image is sent via facsimile over telephone lines. Even at high data rates, a page would take over 13 minutes to send if uncoded, whereas a coded page typically would take less than a minute.

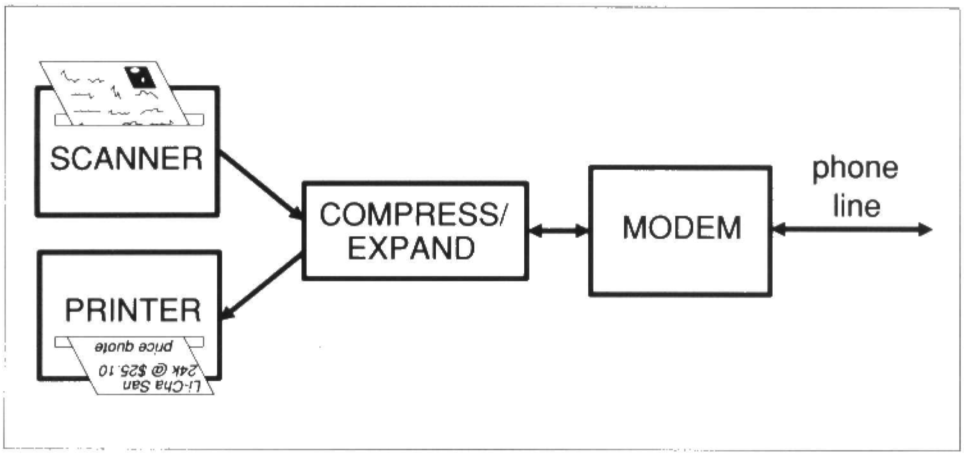


Fig. 1. Simple facsimile machine.

Compression and expansion without information loss

A scanned image consists of horizontal scan lines containing runs of black pixels followed by runs of white pixels. A blank line would consist of all white pixels. The scanner represents black and white pixels with ones and zeros.

If we represent each 'run length' of pixels of a given colour with the number of pixels in the run, we can store the image in much less memory. For example, if 528 white pixels are followed by 92 black pixels, this could be represented by 528 (white) plus 92 (black). Storing these two numbers representing the scan line contents takes much less memory than storing the 620 pixels individually as ones and zeros.

Statistical analysis of typical office and engineering documents shows that certain run lengths occur more frequently than others. So, instead of run length codes, Modified Huffman (MH) coding is used, where shorter coded words represent more frequently occurring run

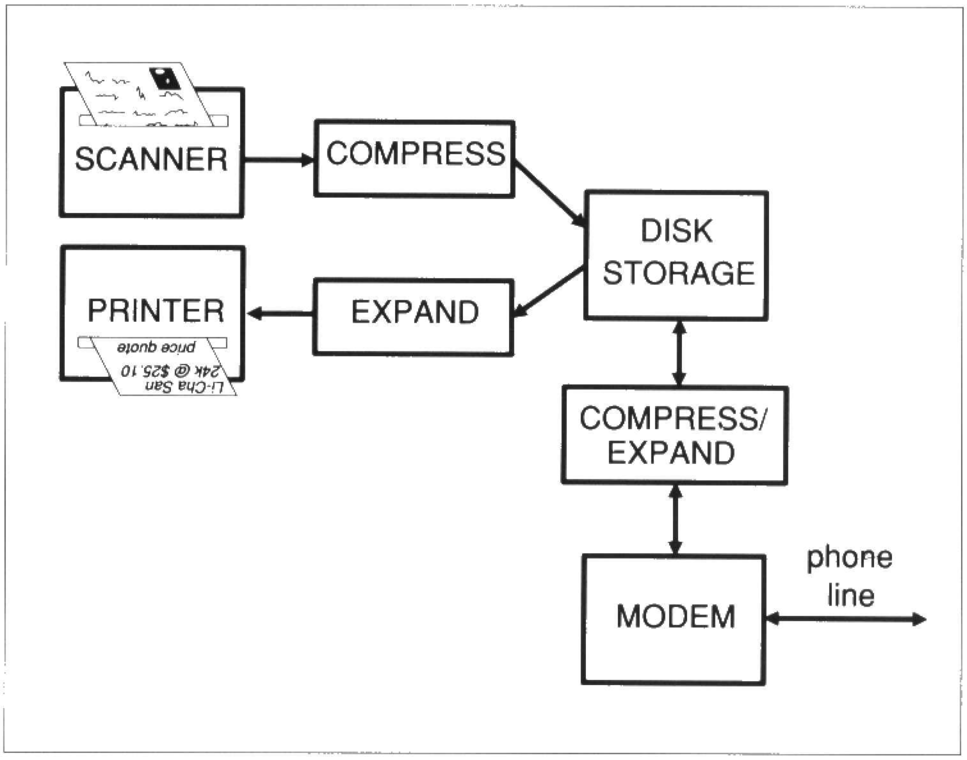


Fig. 2. Store-and feedforward system.

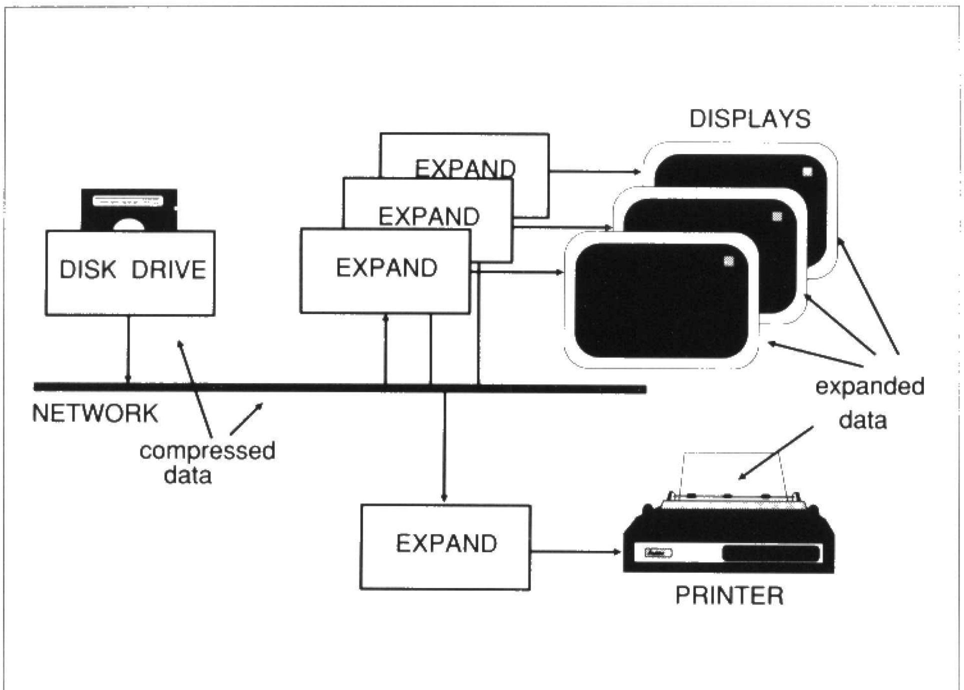


Fig. 3. Typical document storage/retrieval system.

AM95C71 VCEP

- Throughput exceeding an average rate of 50 MB/s when compressing or expanding
- Full CCITT Group 3 and 4 compression/expansion: allows MH, MR and MMR coding and transparent mode
- Dual-bus architecture with single-bus mode option
- Supports bit-boundary image width up to 8191 pixels in one-dimensional (1D) mode and 6911 pixels in two-dimensional (2D) mode
- Has on-chip 6911 pixel reference-line buffer allowing high performance 2D coding
- Provides error detection and recovery capability
- Supports programmable *k*-parameter for 2D coding
- 16-word FIFO on input and output
- Half-duplex operation

lengths. When an image is MH-coded (*compression*) it requires less storage space than run length coded data, and typically yields a file size which is compressed 10 to 15 times versus the raw data. However, the original image can be recovered easily and without loss (*expansion*).

Further analysis of typical scanned document images reveals that about half the number of scan lines are identical to the previous line, and that a further quarter differ by only one pixel. This additional redundancy is removed by using two-dimensional coding, where the current scan line is coded with respect to the previous line. Standard 2-D techniques include Modified Read (MR) and Modified MR (MMR), which typically yield compression ratios (original file size compared to coded file size) of more than 25:1.

Two-dimensional coding is much more sensitive to transmission data errors and requires error protection, while MH coding (one-dimensional) prevents error propagation throughout the document by confining errors to the scan lines in which they occur.

Applications

Image compression and expansion can be used in any system handling binary (black-or-white) images. Such systems include facsimile machines, which typically send and receive documents over comparatively low data rate telephone lines; document storage/retrieval systems, which handle document images, such as office and engineering records or library books, communicate over short, high-speed networks and require quick image access times; and copiers and printers, where an image may be stored internally or received via a network, before being printed.

Facsimile

The CCITT, the U.N.'s telecommunications standards group, developed MH, MR and MMR coding to standardize document facsimile. All of today's fax machines use CCITT standards. A minimally configured fax system is shown if Fig. 1. This will either scan, compress and transmit a document, or receive, expand or print a document.

Fax machines designed today often do more than simply send and receive documents, as shown in Fig. 2. Such machines contain a disk drive and are capable of storing scanned documents for bulk transmission when phone rates are cheaper. In addition, the machine can automatically send the same fax to many remote locations without requiring the page to be scanned each time.

These *store-and feedforward* machines also may contain a 'password' feature, where received faxes will not be printed out or displayed until the correct password is entered. Store-and-feedforward machines compress and expand documents between the user device (scanner or printer) and the disk, or between the disk and the telephone line. These multiple paths in practice require simultaneous compression and expansion to occur within the machine. System cost and complexity can be reduced significantly if the VLSI device handling compression and expansion can deal with both tasks at the same time (full-duplex operation).

Fax transmission speed is usually limited by phone line bandwidth. In an ISDN (*Integrated Systems Digital Network*) environment, where the data rate is 64 kB/s, the AM7971A CEP (*Compression/Expansion Processor*) is still fast enough

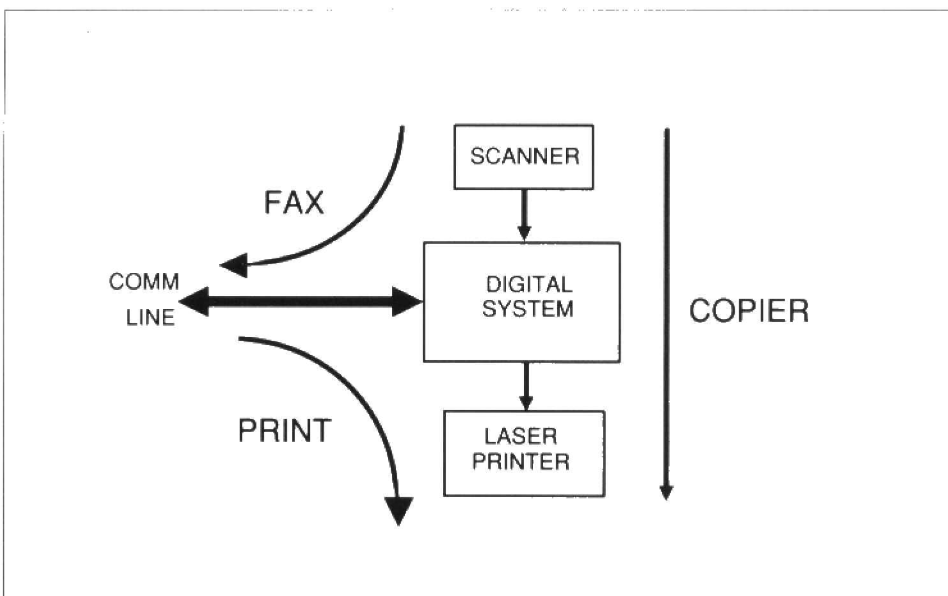


Fig. 4. The intelligent copier/printer is a multi-function system.

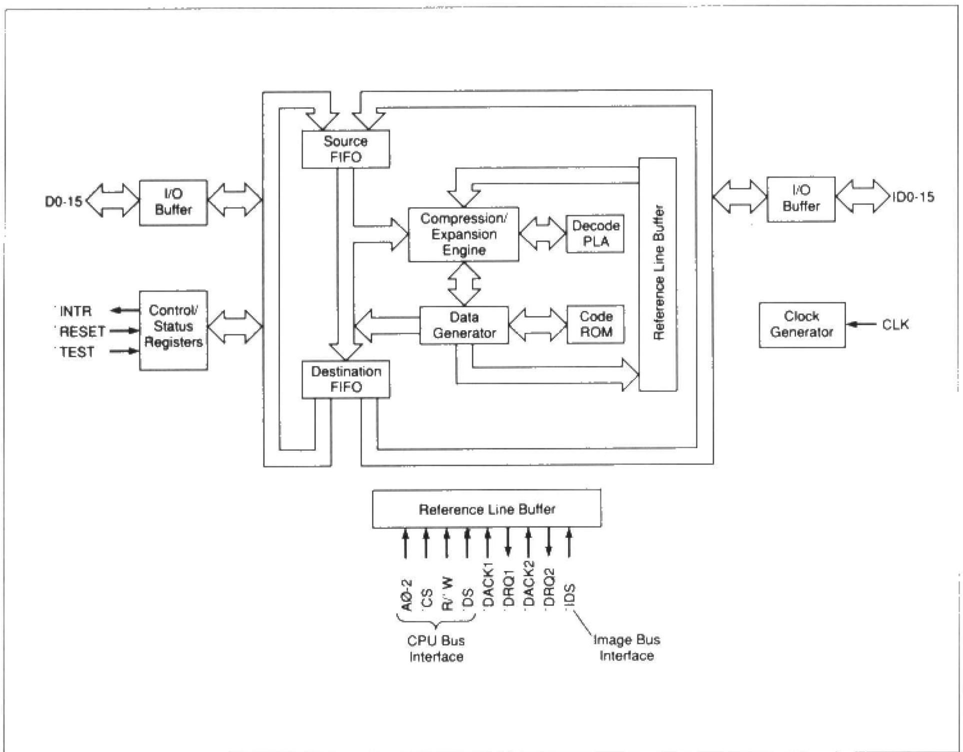


Fig. 5. Block diagram of the AM9571 VCEP.

to allow full use of available bandwidth, even in full-duplex operation. The device has been superseded, however, by the much faster AM95C71 VCEP.

Document storage/retrieval (DS/R)
A document storage/retrieval system can scan, compress and store documents to be retrieved, edited or printed (Fig. 3). Compression speed is important, particularly when a high-speed scanner is used and many documents are input. Expansion speed is often more important, since a user may need to browse through numerous documents before finding the required one. This ideally requires a system which expands a document in human reaction time (the time taken to examine the screen and select the next screen), allowing the user to browse as fast as is desired.

For typical reaction times, document scan resolutions and page sizes, an expansion rate of between 30 and 60 MB/s is required. In addition, since DS/R systems have many different configuration options and speed requirements, compression/expansion VLSI must be flexible in interfacing to a wide range of systems.

The AM95C71 VCEP (*video compression/expansion processor*) can compress and expand even complex documents at rates significantly above 50 MB/s in all coding modes (MH, MR, MMR), while remaining simple, flexible and economical to use.

Intelligent copier/printer (IC/P)
The intelligent copier/printer is a fully electronic, distributed system capable of scanning, printing, sending and receiving documents. It can therefore function as a simple scanner, a networked printer, a facsimile machine or a copier (Fig. 4). Each component in the system performs either compression or expansion: electronic collating is here!

When the system functions as a copier, documents may be scanned, compressed and stored before being expanded and printed. If many copies of the same page are needed, the page need only be scanned once. For multiple copies of a many-paged document, the IC/P allows documents to be collated electronically, eliminating bulky mechanical collators.

In an efficient system, document compression and expansion times must be small compared with scanning and printing times, particularly when electronic collating is done. Copiers need the compression and expansion performance only the VLSI can deliver in integrated form.

AM95C71 Video Compression/Expansion Processor (VCEP)

The AM95C71 VCEP provides the fastest compression and expansion available in integrated form. Most images can be throughput in excess of 50 MB/s — that means over 60 pages per second! Yet for all its speed, the VCEP is economical. By providing a simple user interface together with easy programmability, the device provides speed without unnecessary extras that increase device cost and system development time. The VCEP is 100% CCITT Group 3 and 4 compatible, and is implemented in low-power CMOS technology.

Features of the AM95C71 VCEP include:

- No-loss image compression and expansion
- Fully CCITT Group 3 and 4 compatible coding
- Simple hardware interface; 16 word FIFOs on input and output
- Simple but flexible software interface; only 6 programmable registers
- Bit-alignment capability
- Single or dual-bus operation
- Low-power CMOS
- 68-pin PLCC package

Throughput rates vary depending on image complexity and coding choice. Running at the maximum clock rate of 20 MHz, the AM95C71 typically handles MMR coding at 65–75 MB/s, MR coding

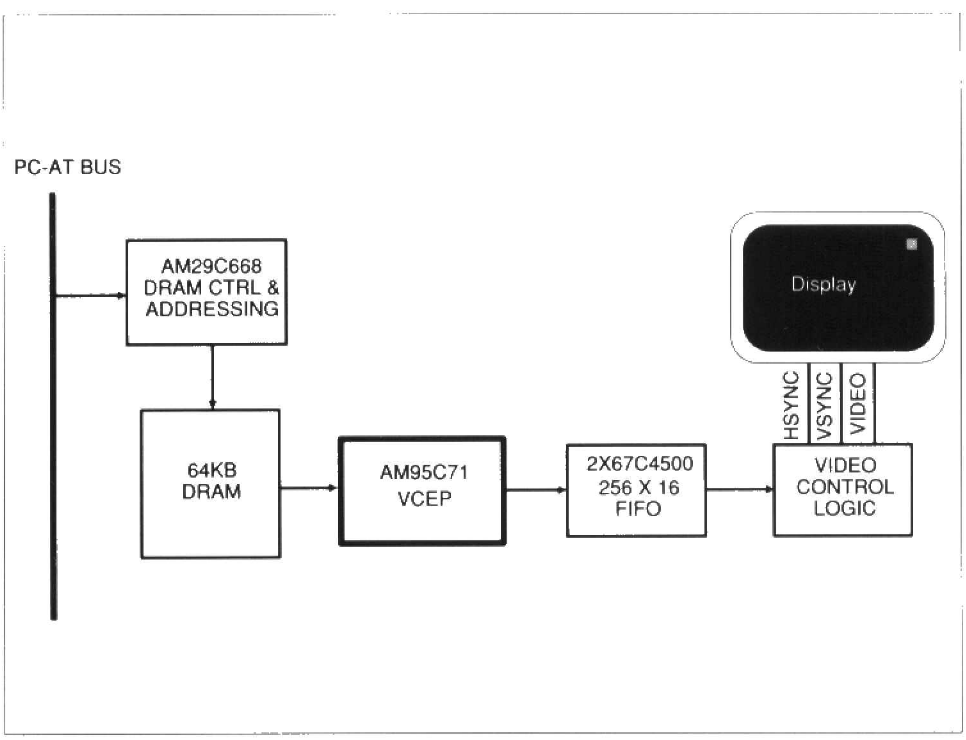


Fig. 6. Simple PC add-on board for document retrieval and display.

(k=4) at 70–80-MB/s, and MH coding at 75–85 MB/s.

The VCEP block diagram is shown in Fig. 5, while typical system block diagrams are shown in Figs. 6 and 7. Note that while implementing the VCEP in a dual-bus configuration gives optimum performance in all cases, a single-bus configuration may be used, sometimes without reduced throughput, depending on system bus traffic.

Further information on the AM95C71 VCEP may be obtained from **Advanced Micro Devices, Inc.** • P.O. Box 3453 • Sunnyvale CA 94088 • USA.

AMD UK can be contacted at (0925) 828008 (Manchester area) or (04862) 22121 (London area).

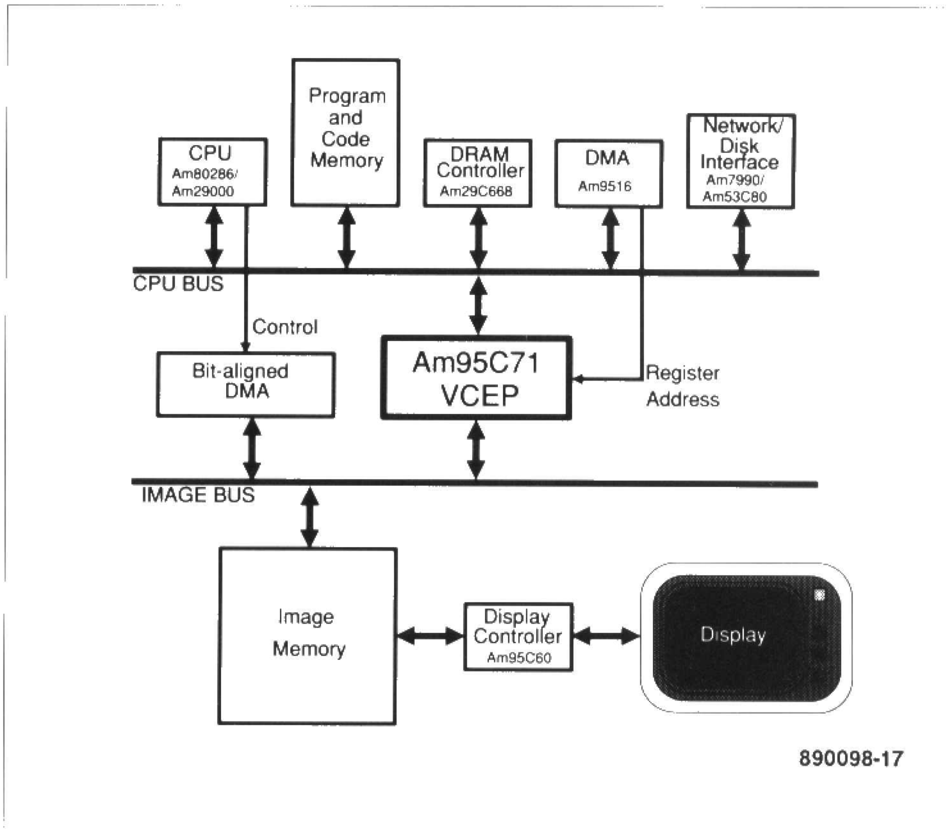
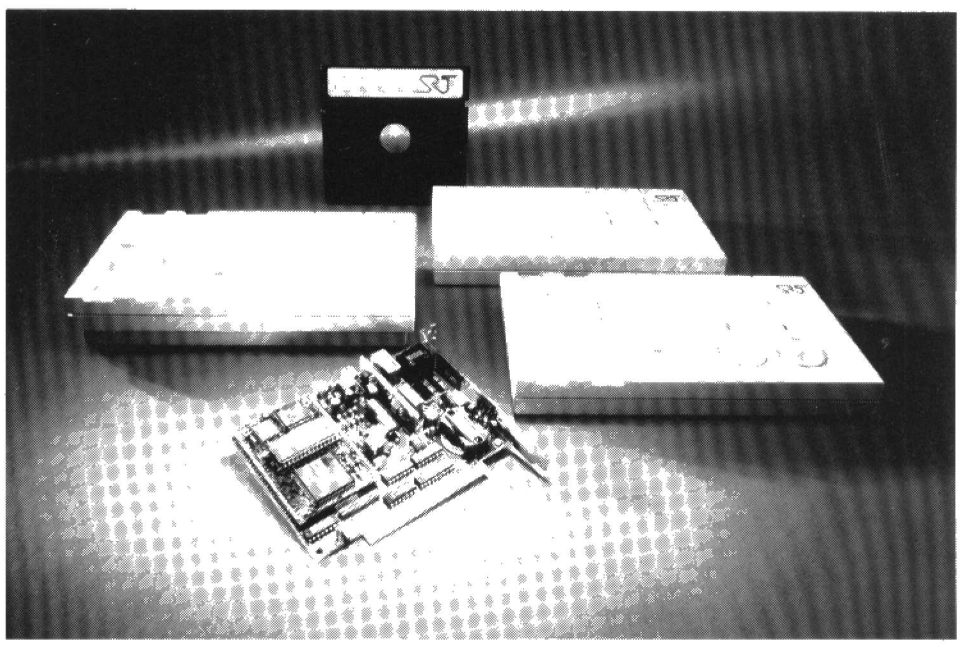


Fig. 7. Typical full-performance VCEP system.

NEW PRODUCTS

New range of modems



Standard Radio & Telefon AB (SRT), Sweden's largest dial-up modem manufacturer, has recently launched a new series of modems.

In contrast to virtually all competitors on the modem market, SRT has developed its own VLSI circuits. The company has been doing this for ten years, and claims that this approach ensures the best possible quality, performance and, above all, reliability.

Apart from the standard-sized modems in the news series, a module is available which contains the same advanced circuits together with a control processor. The module allows other modem and data terminal manufacturers to use a ready-made 'data pump' in their own products.

One of SRT's new modems, the UNILINE, is claimed as one of the first universal modems for IBM-compatible computers. It is available either as a stand-alone unit or as a plug-in card, and is said to meet both European CCITT and North American Bell recommendations, while providing PC communication speeds from 75 to 2400 bits/s, full duplex, with automatic identification and baud rate selection. The Uniline is designed to cope with poor telephone line quality. A communications software package, INDEX, is included with the modem.

Other modems in the series are the simpler BASELINE 2.4 with 2400 bits/s transmission speed, full duplex operation, and the OEMLINE, intended for building into other manufacturers' equipment.

The SRT CRYPTOLINE 2.4 is a desktop modem which protects confidential information with the aid of an internal encryption procedure which is very difficult to decipher or disable.

HYPERLINK is another new item. This terminal system enables simultaneous speech and data communication over existing internal or external telephone lines. The system allows the connection of four terminals plus a telephone set, using one and the same two-wire link. HYPERLINK operates at speeds of 4×9600 bits/s or 2×19200 bits/s to allow a local network to be built up very easily without the need to install new wiring.

Standard Radio & Telefon AB Box 501 S-16215 Vällingby SWEDEN. Telephone: +46 8 7394000. Fax: +46 8 7394147.

MULTI-LAYER PCBS

by A.J. Kool (ULTimate Technology, Norcross, USA)

The introduction of multi-layer printed-circuit boards challenged designers in their creativity to use these additional layers efficiently. One of the ways to increase the area left for actual wiring is the use of buried vias. Buried vias connect copper layers in layer pairs. They are not drilled through the entire board, leaving more space for routeing on the other layers. Since the advent of surface mount technology – SMT – these buried vias have become more popular with designers. The increased use of these vias, however, presents the designers with a new set of problems that only few CAD systems are able of coping with.

What are buried and blind vias?

The idea of creating buried and blind vias comes from the manufacturing of the printed-circuit board. A multi-layer board is built from a set of (thin) double-sided boards. The copper layers of these thin boards form the copper layers of a multi-layer board. They are all etched and then stacked together, separated by insulation layers. This is the lamination process illustrated in Fig. 1.

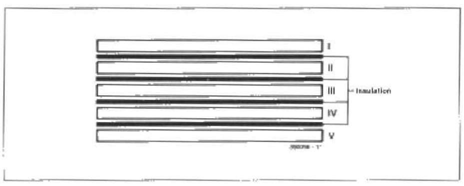


Fig. 1. The layers are laminated together as thin boards.

Buried vias are created by drilling and metallization of the thin boards before the lamination process. By doing this, the layers of the thin boards are connected with vias. At this point, the whole set of thin boards is laminated, the vias in the middle layers are buried (not visibile from the outside) and the vias in outer layers are blind (visible from one side only) – see Fig. 2. For the designers, there is no difference between buried and blind vias.

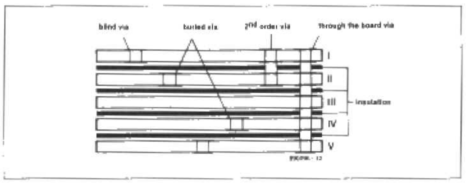


Fig. 2. Buried and blind vias are located in the board.

When the idea of drilling and metallization is taken one step further, we may laminate boards I and II, drill and metallize the resulting board and complete the rest of the lamination process – see Fig. 3.

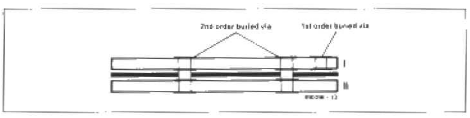


Fig. 3. Layers I and II are drilled and metallized before laminating the other layers.

Vias created like this are called 2nd order buried vias. By using 2nd, 3rd or higher order vias, we obtain a complicated lamination and metallization sequence that increases the routeability of printed-circuit boards, but also their cost.

Using a CAD system

Until recently, CAD tools would only handle the buried vias between layer pairs. No provisions were available to have vias through more than two layers. Now SMT is used more extensively, more designers are looking for CAD tools that handle buried and blind vias in a highly automated way. To have a CAD system automatically handle these vias, it must first know how the designer is planning to laminate the PCB. Are 2nd, 3rd and higher order buried vias allowed? These are the things a CAD system must know before vias can be used. When this is known by the system, a via placed between the TOP layer and Inner 3 can be computed to be 'through the board' or a 2nd order (if this was possible in the lamination sequence).

Another aspect the CAD system must deal with is the size of the drill hole. A multi-layer board with 10 layers is not (much) thicker than a 2-layer board. A single layer-pair of the 10-layer composite board is very thin. The via drill diameter may be smaller for these thin boards than for the through-the-board vias. The CAD system must compute the number of layers the drilling hole will cross and select the drill diameter accordingly.

If the designer wants to extend 1st order buried vias to a higher order, the drill diameter may need to be changed. To be able to manufacture the board, the PCB manufacturer needs to have a drill file for each layer-pair and each half-product that has buried vias. Finally, a drill file for the complete board is needed for the through-the-board vias and component pins. The 10-layer board of Fig. 4 needs 8 drill files: 5 for the layer-pairs, 2 for the half-products and 1 for the complete board.

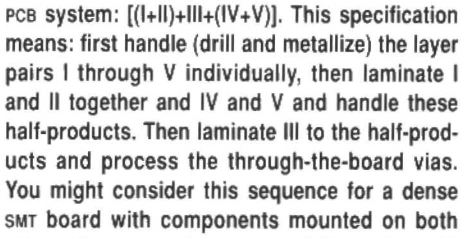


Fig. 4. Example of a lamination sequence specification for a 10-layer board in the ULTiboard PCB system: [(I+II)+III+(IV+V)]. This specification means: first handle (drill and metallize) the layer pairs I through V individually, then laminate I and II together and IV and V and handle these half-products. Then laminate III to the half-products and process the through-the-board vias. You might consider this sequence for a dense SMT board with components mounted on both sides. The layers of layer-pair III (inner 4 and 5) would be the power and ground planes. The others (TOP, INNER 1, 2, 3, 6, 7, 8 and BOTTOM) will be signal layers.

Conclusion

Currently available CAD systems are sufficient for handling boards with 4 signal layers. In a few years' time, when the need arises for a board with 6 or more layers, all CAD systems will almost certainly have to be able to support the higher order buried vias to some extent.

DC-AC POWER CONVERTER

J. Ruffell

Holidaymakers, do not forget this low-cost power converter when you are packing for this year's camping tour. The converter works from the car battery, is simple-to-build from standard components, and provides you with up to sixty watts to power mains-operated loads such as a shaver, a small fluorescent tube, and (dare we suggest it?) a soldering iron.

There is nothing to beat the good old camp fire, candles, or pale moonlight to light holidaymakers gathered for the evening barbecue on the camping site. An electric light source, on the other hand, is the thing you want when tentpegs are to be driven in the ground in the middle of the night, somewhere, in the dark, on an unfamiliar site. An AC converter is also handy for the daily shave, for the portable TV set, a small fluorescent tube, a radio, oscilloscope or computer.

Circuit description

The circuit diagram of Fig. 1 shows that the DC-AC power converter is built from commonly available and inexpensive parts. Circuit IC$_1$, a CMOS Type CD4047, is used as an astable multivibrator whose outputs, Q and $\overline{Q}$, supply a square wave signal that has a frequency of about 50 Hz. To prevent excessive loading of the chip outputs, the complementary signals are fed to the gates of Type BS170 low-power MOSFETs. These transistors are capable of switching at high speed, they guarantee low turn-on and turn-off times, and provide sufficient drive for the bipolar power stage composed of drivers T_3-T_4 and power devices T_5-T_6. Like the MOSFETs, the transistors used in the power stage are selected for their switching speed, with an aim to keep dissipation in T_5 and T_6 as low as possible. Zener diodes D_2 and D_3 protect the power transistors against voltage peaks generated by the transformer, which forms an inductive load.

The power transformer is a standard type, i.e., not a toroid, and is used 'the other way around' to step up the low voltage to the mains voltage. The low-voltage winding with its centre tap forms the primary. The centre tap is not connected to ground as usual in most power supplies, but to +12 V. The power transistors, T_5 and T_6, alternately take the outer connections of the primary to ground, passing considerable current. This, in turn, induces a voltage in the secondary, which is the mains winding in this case. A fuse completes the AC-DC converter.

50 Hz quartz-controlled

Most radio alarm clocks use the frequency of the mains voltage as the timing reference. A small extension circuit enables the DC-AC converter to supply the mains voltage at a constant and accurately defined frequency of 50 Hz. The printed-circuit board of the converter is provided with a connection for accepting the 50 Hz signal.

Figure 2 shows that the 50 Hz reference signal is derived from a 3.2768 MHz quartz crystal. The circuit uses only two CMOS ICs, and operates from 12 V. The quartz crystal is an inexpensive type commonly used in clock timebase circuits. The frequency of oscillation is trimmed with C_2. A fixed, ceramic, 12 pF capacitor may be preferred over the trimmer in some cases, and results in a frequency deviation that is perfectly acceptable for the application in question. When the trimmer is used, it is adjusted for a frequency of 204.8 kHz at test point TP. The 50 Hz signal available at point 3 of the timebase is connected to point 2 of the DC-AC converter. In this configuration, a wire link is installed between points 5 and 6. When

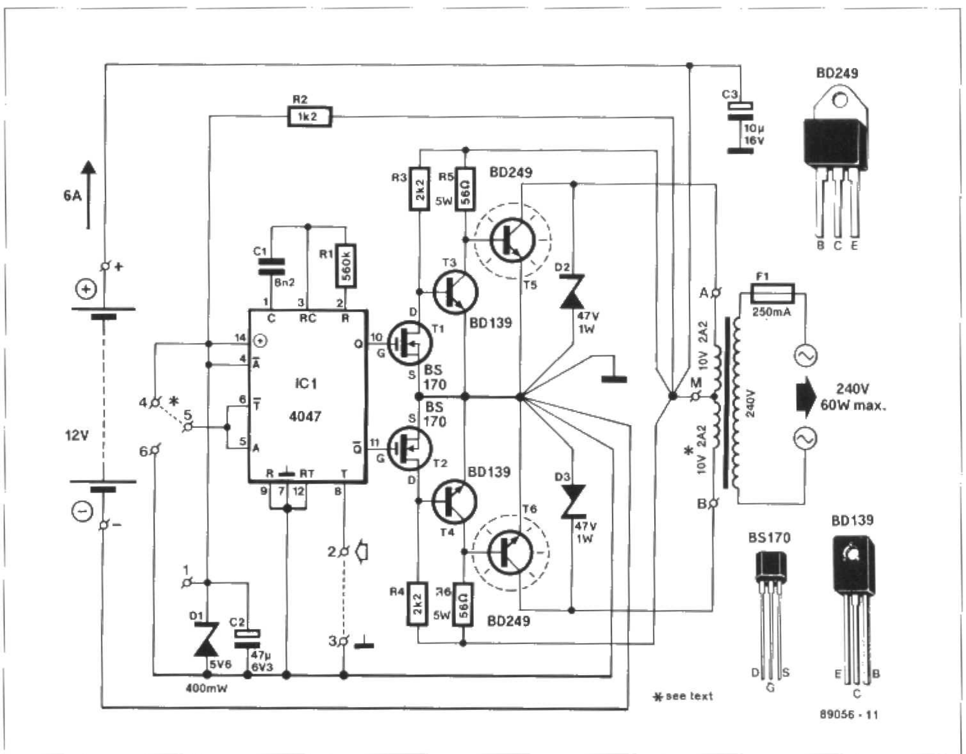


Fig. 1. Circuit diagram of the DC-AC power converter. The wire links enable the use of an optional 50 Hz timebase.

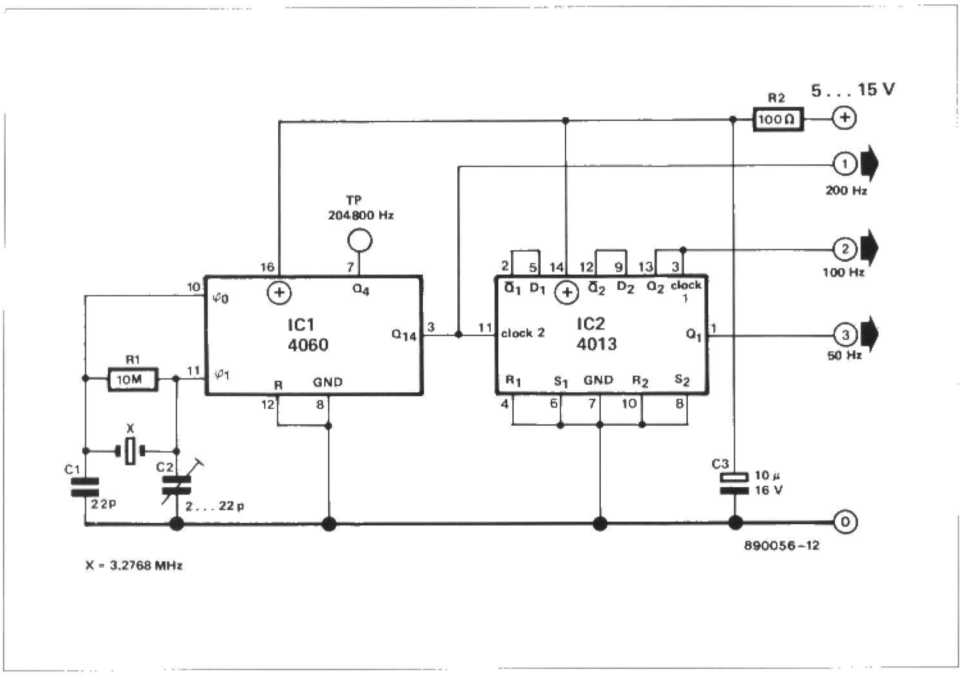
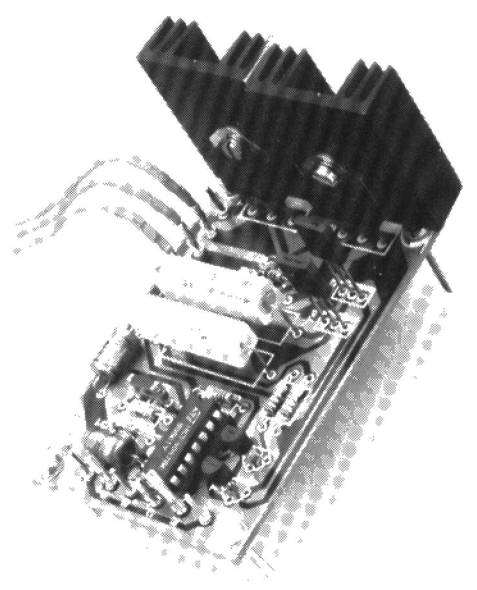


Fig. 2. This quartz-crystal controlled timebase is an optional extension of the power converter.



the external timebase is not used, the wire link is installed between points 4 and 5, while point 2 is grounded via a link to point 3.

Construction and practical use

There is little to say about the construction of the power converter because the population of the printed circuit board (Fig. 3) is entirely straightforward. By virtue of the high overall efficiency, the power transistors can do with a relatively small heat-sink. When the unit is mounted in a metal enclosure, the transistors are conveniently bolted on to a side panel. Do not forget to use insulating washers and a touch of heat-conducting compound. The introductory photograph shows the prototype in a sturdy metal enclosure with a shaver-type output socket and heavy-duty wander sockets for connecting the battery cable.

The low-voltage winding of the transformer is switched to achieve high efficiency. As a result, the generated high voltage is a fairly clean square wave, which remains largely rectangular with

Parts list

Resistors:
R_1 = 560k
R_2 = 1k2
R_3;R_4 = 2k2
R_6;R_7 = 56Ω; 5 W

Capacitors:
C_1 = 8n2
C_2 = 47µ; 6V3
C_3 = 10µ; 16 V

Semiconductors:
D_1 = zener diode 5V6; 400 mW
D_2;D_3 = zener diode 47 V; 1 W
T_1;T_2 = BS170 (Cricklewood Electronics)
T_3;T_4 = BD139
T_5;T_6 = BD249
IC_1 = 4047

Miscellaneous:
F_1 = fuse 100 mA; slow.
Tr_1 = mains transformer 2×10 V; 2.2 A.
Heat-sink for T_5 and T_6; max. 4°C/W.
Insulating washers and bolts for T_5 and T_6.
PCB Type 890056 (not available through the Readers Services).

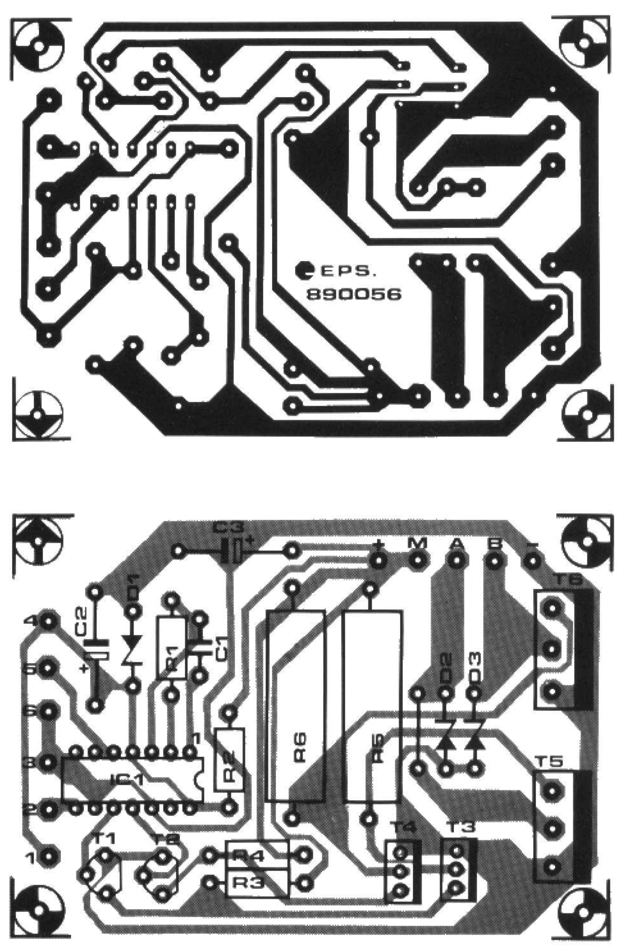


Fig. 3. Printed-circuit board for the DC-AC power converter.

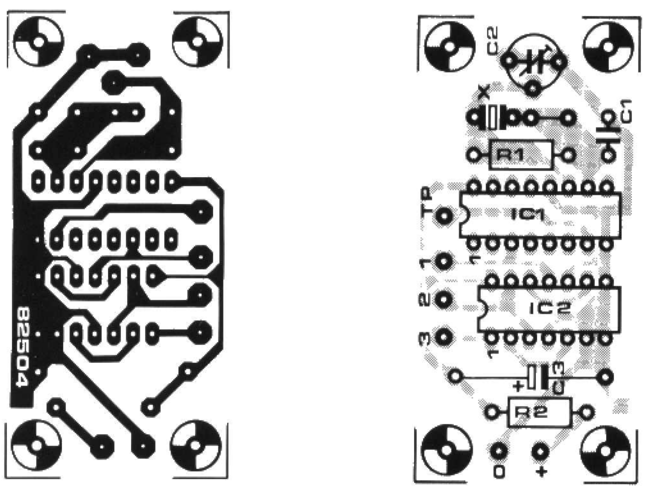


Fig. 4. Printed-circuit board for the optional timebase circuit.

Parts list

Resistors (±5%):
R1 = 10M
R2 = 100Ω

Capacitors:
C1 = 22p
C2 = 22p foil trimmer (green)
C3 = 10µ; 16 V

Semiconductors:
IC1 = 4060
IC2 = 4013

Miscellaneous:
X1 = 3.2768 MHz quartz crystal.
PCB Type 82504 (not available through the Readers Services).

some overshoot even when an inductive load, such as a mains transformer, is powered.

Regulation of the output voltage is purposely not implemented to keep the overall cost of the converter as low as possible. This means that the open-circuit output voltage is higher than the loaded output voltage, an effect which is caused mainly by the copper losses of the transformer windings.

It will be evident that the output voltage is also dependent on the battery voltage. At a battery voltage of 14 V, the output voltage is about 10% higher than at 12 V. When the power converter is frequently used with relatively heavy loads (between 40 and 60 W), such as a portable TV, it is recommended to use a 2×9 V transformer rather than a 2×10 V type.

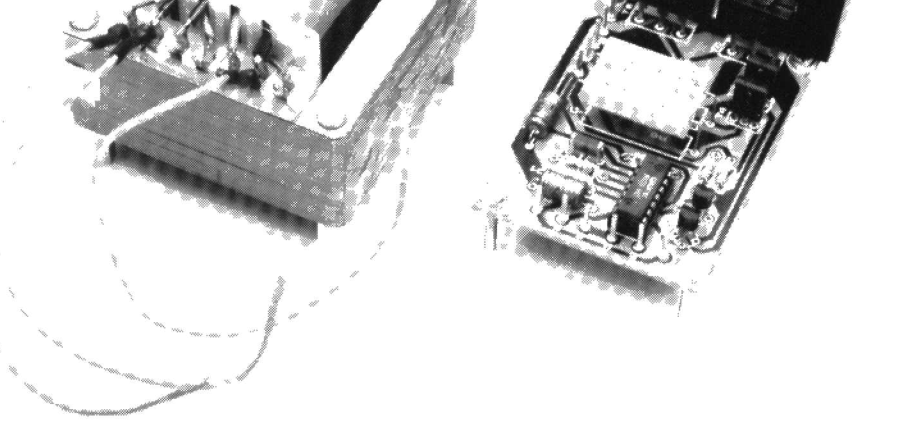
At an input voltage of 12.0 V, and loaded with a light bulb, the prototype gave the following results (the transformer was a 240 V/2×10 V type):

Bulb wattage	U_o (V_{rms})
0	261.5
25	242.1
40	222.2
75	187.2
100	165

The power converter gave no problems when used to power the ohmic-inductive load formed by an electric shaver. Encouraged by this result, we connected a Commodore C64 computer plus monitor (40 VA). Although the output voltage dropped to about 210 V, the computer worked all right. In the case of equipment not functioning owing to a too low voltage, it is best to resort to the use of a 2×9 V transformer in the converter.

An experiment with a cassette recorder deck gave less satisfactory results: the heads picked op harmonics of the converter's output signal, so that the equipment produced unacceptable background noise. The same cassette deck was known to be sensitive in this respect, however, since it produced almost the same background noise when installed near the light dimmer in the living room at home.

In many cases, it is advisable to use a mains filter between the power converter and the load when this is mainly inductive and rated at more than about 30 W. Such a filter may be purchased ready-made, but a common suppressor choke as used in dimmer circuits, in combination with a 470 nF/400 V capacitor will also do the job.



INTERMEDIATE PROJECT

A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available parts.

3c. Function generator, final part

J. Bareford

The low-cost function generator we set out to describe three months ago is completed with a waveform selection circuit, an output amplifier and a power supply.

You are almost there if the modules described in the previous two instalments of this article (Refs. 1 and 2) have been built and tested. For those eager to finish the project we have designed two more modules that are required to give the test instrument all the functions users of a function generator have come to expect.

The rectangular and triangular wave generators described earlier are, in principle, modules that do not require additional circuitry to enable their use in quite a few applications. There is, however, a good reason for adding an amplifier module: both generators have a relatively high output impedance, so that the output voltage drops to an unacceptable level when the generator output is loaded with a relatively low resistance or impedance. A buffer amplifier prevents this by acting as an impedance converter between the generator output and the load. All that is required for this function is, in principle, supply current. Hence, a mains power supply is discussed also.

Equal levels

Those of you who have closely read Part 1 of this article, and those who have already managed to produce a working rectangular/triangular wave generator, will have come to the conclusion that the peak-to-peak voltage of the rectangular signal is much greater than that of the triangular signal. The effect is caused by the comparator, whose output voltage swings from +8 V to −8 V when the triangular voltage exceeds −1.7 V, or drops below +1.7 V (in practice, these levels may be a little lower). This results in a rectangular voltage of 13 V_{pp}, while the amplitude of the triangular voltage is only 3 V_{pp}. Because it is derived from the triangular signal, the sine-wave has an even lower level of about 2 V_{pp}.

Owing to the way they are generated and derived, the three generated voltages differ both as regards shape and amplitude, so that some means will have to be found to ensure that the generator supplies a constant, adjustable, output amplitude, irrespective of the selected waveform.

Three voltage dividers, one for each waveform, and dimensioned to supply equal output amplitudes, form a suitable starting-point for solving the problem of unequal amplitudes. The basic circuit of Fig. 1 follows a slightly different approach by using three series resistors and one potentiometer common to these. P_1 has on it the output voltage, and is the second resistor for all three voltage dividers. The 10 kΩ and 1 kΩ resistors marked with an asterisk are soldered direct to the relevant contacts of the waveform selection switch, S_1. The 220 Ω resistor replaces R_3 in the sine-wave converter. The 1 nF capacitor is C_3 in the sine-wave converter.

Power opamp

The signal voltage on P_1 is the same for all three waveforms, and a part of it is fed to an output amplifier via the wiper of the potentiometer. The full circuit diagram of the buffer amplifier, given in Fig. 2, shows an apparently standard application of an operational amplifier. The cir-

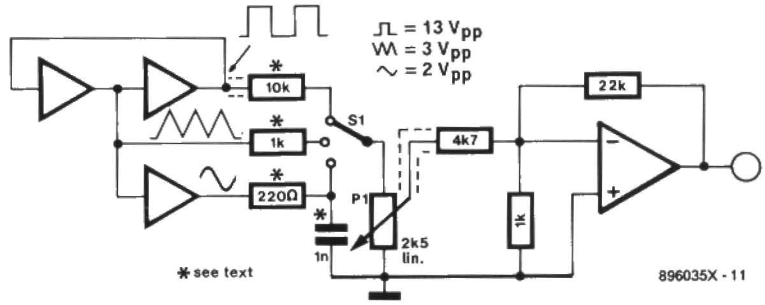


Fig. 1. Component configuration around the waveform selection switch.

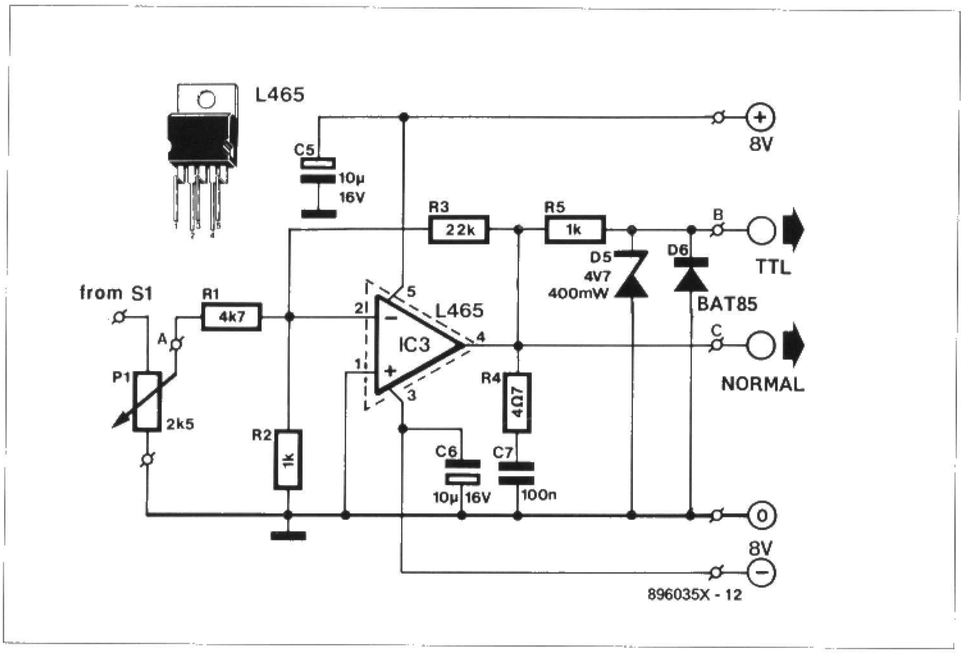


Fig. 2. Output amplifier set up around the Type L465 power operational amplifier.

cuit is fairly unusual, however, because it is based on a power operational amplifier Type L465. The datasheets of this device from SGS state a maximum output current of no less than 3 A, which makes the L465 a quite remarkable component. Fortunately, this does not mean that it can not be used as an ordinary, low-power, opamp.

In practice, some characteristics of the L465 require special attention to be paid to the design of the buffer amplifier. One of these is that the overall gain must not be allowed to become smaller than 20. Below this limit, the bandwidth of the L465 is so large that oscillation occurs readily. The gain-bandwidth product of an opamp is a fairly constant figure given in the datasheets. The gain often rises, therefore, to uncontrollable levels when the bandwidth is not limited at the low side. The effect is caused by internal capacitances of the opamp, and stray capacitance in the circuit, which results in signals of a relatively high frequency being fed back more readily than low-frequency signals.

In theory, therefore, it must be assured that the gain can not drop below 20. In the present case this is still far too high, however, because the function generator supplies more output voltage than could possibly be handled by the buffer amplifier.

Figure 1 shows that the output voltage of the rectangular wave generator is 13 V_{PP}. This value is reduced to 2.6 V_{PP} by the 10 kΩ resistor and P_1:

$$(2.5\ k\Omega \times 12.5\ k\Omega) \times 13\ V_{PP} = 2.6\ V_{PP}.$$

In the theoretical case of the amplifier providing a gain of 20, and P_1 being set to maximum output (2.6 V_{PP}), the output voltage would become $20 \times 2.6 = 52\ V_{PP}$ (provided, of course, that the output amplifier is fed from a supply voltage higher than 52 V).

The use of such high supply voltages is inconvenient and therefore out of the question in this case. Besides, the L465 has a maximum supply voltage rating of 40 V. So, the gain must be made smaller than 20. The resistor values in the feedback path (see Fig. 1) result in a gain of approximately 5:

$$22\ k\Omega\ /\ 4.7\ k\Omega = 4.7$$

Oscillation is prevented by R_2, the 1 kΩ resistor between the – input of the opamp and ground. This resistor turns the amplifier into a summing amplifier whose output voltage is determined by the sum of the voltages applied via R_1 and R_2 (see Figs. 1 and 2). Since these resistors have a different value, the signal applied via R_1 is given a different gain than the signal applied via R_2. The gain for the signal applied via R_1 is about 5 as calculated above, while that for the signal applied via R_2 is

$$R_3\ /\ R_2 = 22\ k\Omega\ /\ 1\ k\Omega = 22.$$

The bandwidth is kept within safe limits, and oscillation is prevented, by making the gain of the second signal greater than 20.

But why bother about a second signal at all, when the output amplifier needs to supply only one signal? The answer is fairly simple. Looking closely at the circuit diagram, it will be noticed that there is no such thing as a second signal, because the signal applied to R_2 is....ground! This means that a constant voltage of 0 V is amplified 20 times. In conclusion, R_2 alone limits the bandwidth by making sure that the gain is higher than 20 at all times. The fact that the signal amplitude subject to this gain is nought is irrelevant.

Two amplifier outputs

The output of the power opamp, pin 4, is connected to a *Boucherot-network*, R_4-C_7, which ensures that the amplifier is always loaded. This is particularly important at relatively high frequencies — without the network, the amplifier would oscillate readily in the absence of a load. The network also enables the driving of highly inductive loads.

The Boucherot network does not affect the shape of the rectangular, sinusoidal and triangular waveforms available at the NORMAL output of the function generator. An additional output has been created for driving TTL (*transistor-transistor logic*) circuits. Components R_5, D_5 and D_6 prevent the output voltage going lower than 0 V (R_5-D_6) and exceeding +5 V (R_5-D_5). The Type BAT85 diode is used in position D_6 because it has a forward drop of only 0.3 V. This means that the lowest drive voltage for TTL circuits is –0.3 V, which is safe for most applications. D_6 must never be replaced by the ubiquitous 1N4148 because this has a much higher forward voltage that may cause damage to TTL

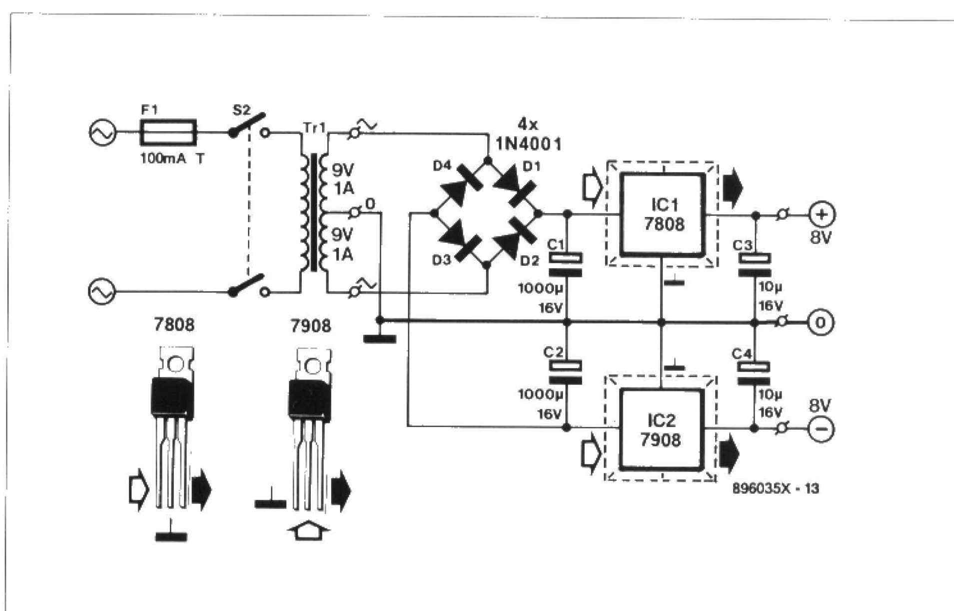


Fig. 3. Circuit diagram of the generator's symmetrical power supply.

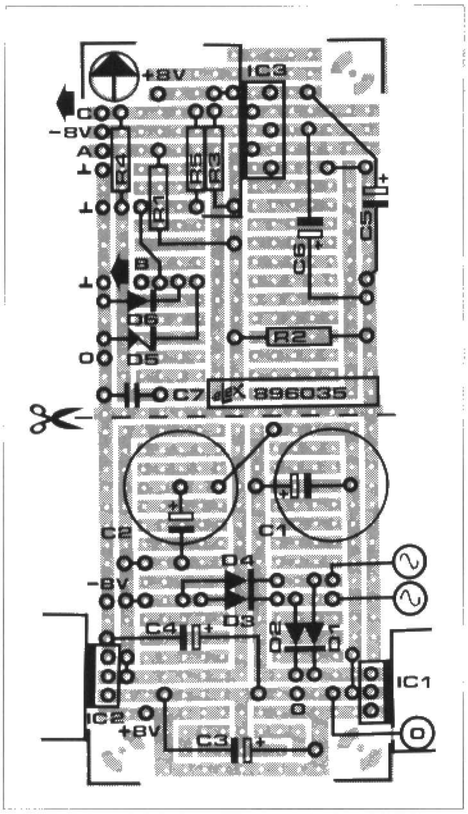


Fig. 4. Component mouting plan of the output amplifier board (upper section), and the power supply board (lower section).

Parts list

Resistors (±5%):
R_1 = 4k7
R_2;R_5 = 1k0
R_3 = 22k
R_4 = 4Ω7
Non-marked resistors (Fig. 1):
10k, 1k0, 220Ω
Capacitors:
C_1;C_2 = 1000μ; 16 V
C_3;C_4;C_5;C_6 = 10μF; 16 V
C_7 = 100n
Semiconductors:
D_1–D_4 = 1N4001
D_5 = zener diode 4V7; 400 mW
D_6 = BAT85 (listed by Cricklewood Electronics)
IC_1 = 7808
IC_2 = 7908
IC_3 = L465 or L165 (SGS; listed by Universal Semiconductor Devices Ltd, and Electro-Mail 0536 204555)
Miscellaneous:
3 heat-sinks TO220-style.
F_1 = 100 mA fuse; slow.
S_1 = 3-way, single-pole rotary switch.
S_2 = double-pole mains switch with indicator light.
Tr_1 = PCB-mount mains transformer; 2×9 V at 1 A.
PCB Type UPBS-1 (see Readers Services page).
Veroboard for mains transformer.
Two-way PCB terminal block.
Screened wire.
ABS enclosure.
2 BNC sockets.
Mains entrance socket with integral fuse-holder.
Knobs.

circuits.

The TTL output is, of course, only used with the generator set to supply a rectangular wave signal, and P_1 set to maximum output level. The TTL output may be used for driving CMOS circuits, but only if these are fed from a 5 V supply. D_5 may be omitted in this case, provided P_1 is set such that the logic high level is equal to the supply voltage of the digital circuit.

Power supply

The use of the proposed output amplifier makes it impossible to power the function generator modules from a pair of 9 V batteries. The mains supply shown in Fig. 3 easily meets the current demand of the generator inclusive of the power amplifier. The symmetrical supply is a standard design based on a mains transformer with a centre tap, a bridge rectifier plus smoothing capacitor, and a pair of three-terminal fixed voltage regulators with associated decoupling capacitors.

One board for two circuits

The component mounting plan in Fig. 4 shows that the power supply and the output amplifier fit readily on a single printed-circuit board Type UPBS-1 (*universal prototyping board size-1*). The board is cut, however, as indicated before any components are fitted. If it is left in one piece, the –8 V supply rail is short-circuited to ground via the track below the dotted line pointed at by the scissors.

Now mount the components on the board, and solder them in place. Pay attention to the polarization of the diodes and the electrolytic capacitors, and the

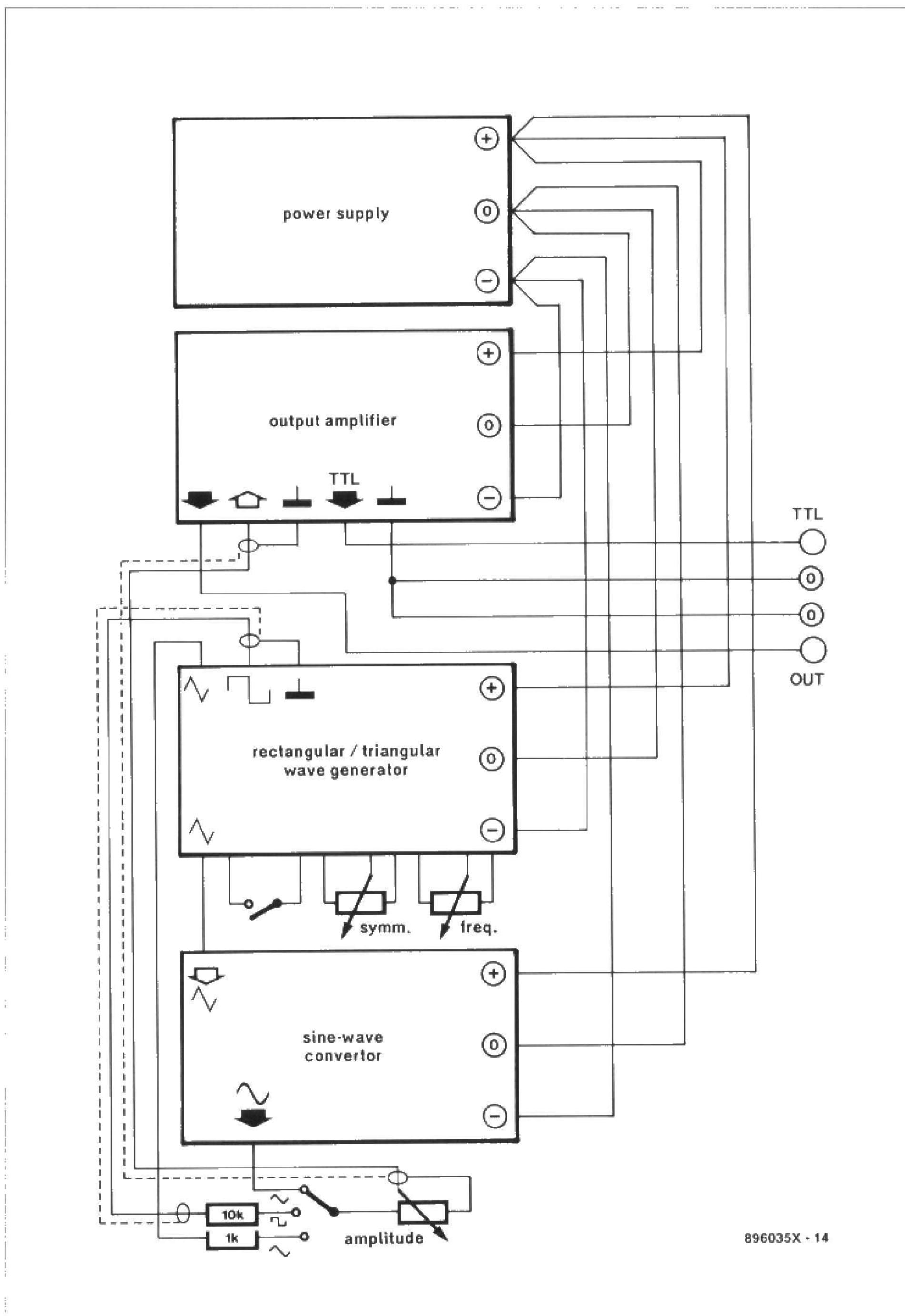


Fig. 5. Wiring diagram of the modular function generator.

orientation of the integrated circuits. Also do not forget the wire links.

As indicated in the circuit diagrams of Fig. 2 and 3, all ICs on the two boards require heat-sinks. A small U-shaped piece of aluminium, or a ready-made TO220-style heat-sink, is adequate for IC_1 and IC_2. A similar heat-sink may be used for the power opamp, IC_3, but care should be taken to prevent short-circuits with the surrounding components, including the wire link near pin 5.

For reasons of safety, the mains transformer is fitted on to a separate circuit board, which may be a piece of stripboard. The mains cable is connected to the transformer's primary winding via a suitable 2-way terminal block. Use a sharp hobby knife and a powerful soldering iron to remove all tracks in between and around those carrying the mains voltage.

Wiring

The completed amplifier and power supply boards are tested and then connected to the other modules in the function generator. The wiring diagram of the complete instrument is given in Fig. 5.

The supply wires to the amplifier, the rectangular/triangular wave generator and the sine-wave converter are connected as separate pairs to the power supply. This is done to prevent the modules interfering with each other's operation via the supply wires. To eliminate any risk of cross-talk and distortion because of the high amplitude of the rectangular signal and its harmonic content, screened cable must be used for two connections. The first is that between the output of the rectangular wave generator and the waveform selection switch, and the second, that between the amplitude control and the amplifier input.

The drawing in Fig. 5 does not show the mains-connected parts, whose wiring is apparent from Fig. 3. Every care should be taken to ensure proper insulation of each and every terminal or wire that is at mains potential. Call in the assistance of an experienced engineer if you have never before worked with circuits operated from the mains.

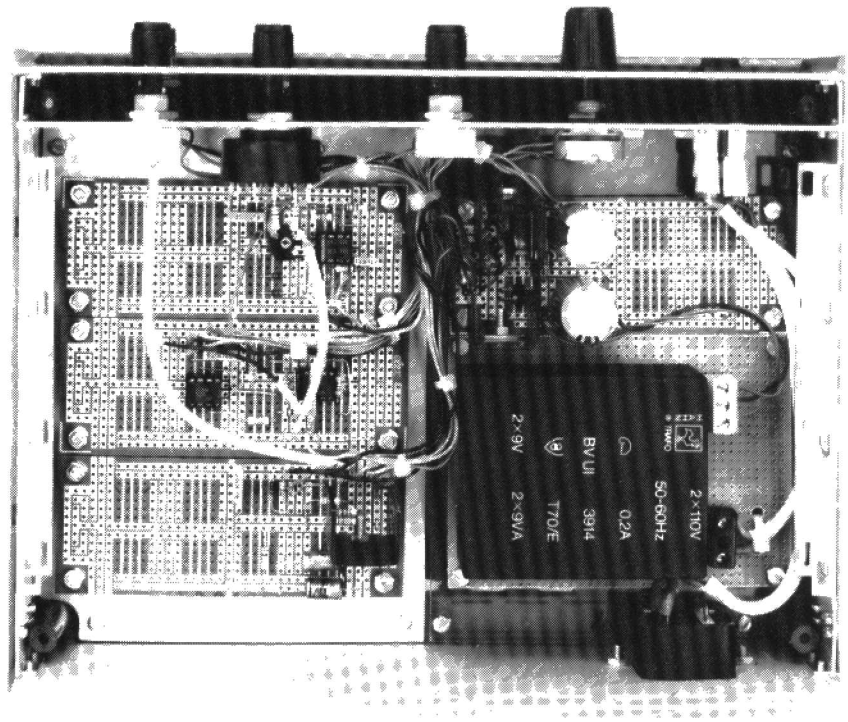


Fig. 6. A look into our completed prototype.

The photograph of Fig. 6 shows a top view of the opened enclosure with all modules installed inside. The two pieces of screened wire and the two resistors soldered direct to S_1 are clearly visible. The prototype has four instead of three boards because the power supply and the output amplifier were built on separate printed circuit boards.

Figure 7 shows a suggested design for the front panel of the function generator. This design is enlarged as required by photocopying, cut to size and secured on to the front panel with a suitable adhesive, and, finally, covered by a protective plastic film.

References:

1. Intermediate Project 3a: Function Generator. *Elektor Electronics* April 1989, pp 53–57.

2. Intermediate Project 3b: Sine-wave converter. *Elektor Electronics* May 1989, pp 14-15.

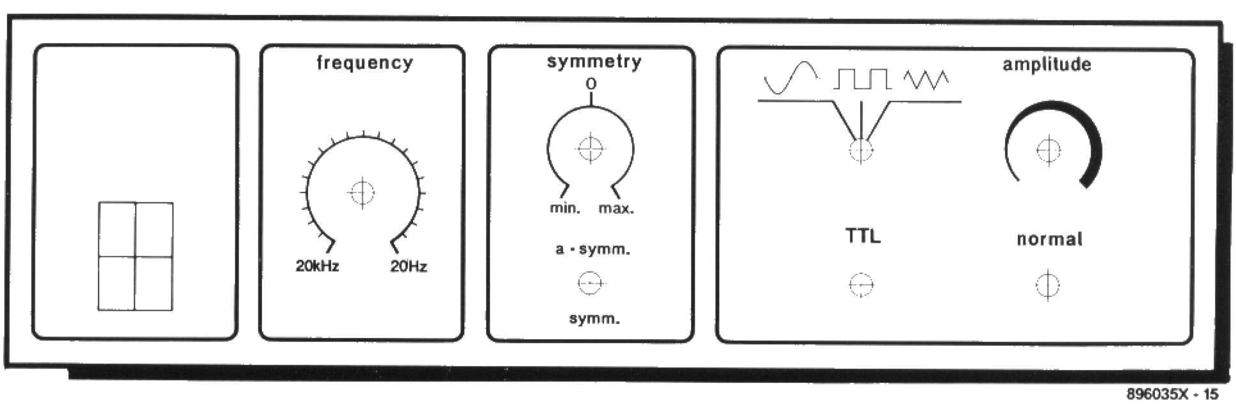


Fig. 7. Suggested front-panel design.

THE DIGITAL MODEL TRAIN – PART 5

by T. Wigmore

This fifth part in the series discusses the Elektor Electronics Digital Train System, which makes possible the independent control via the rails of many locomotives, turnouts (points) and signals. It also makes provision for storing monitoring signals and contains as standard an RS232 interface.

Some mention was made in the previous five articles of digital train control systems with particular reference to the Märklin system and also the Elektor Electronics Digital Train System. From now on the articles will deal specifically with the latter.

The Elektor Electronics system is intended primarily for home construction, which makes it considerably cheaper than proprietary systems, and also gives you the opportunity of making it as simple or as complex as you like. At present, our system offers more facilities and possibilities than any other system we have seen.

The Elektor Electronics system is of modular construction—see Fig. 34. Table 4 compares it with the Märklin system.

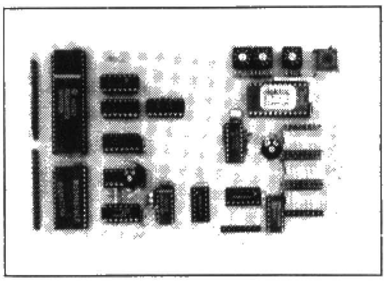
Design considerations

The main PCB of our system plus at least one booster unit (to be published next month) and a power supply form the most basic configuration of the control system. This will accommodate sixteen controllers each consisting of a simple slide potentiometer. These alone will save the home constructor a lot of money, since, for instance, a Märklin controller costs about £60–£70. Multiply that by sixteen and you arrive at a tidy sum. The sixteen slide potentiometers will probably cost you litle more than just one Märklin controller.

Each controller is parallelled by two switches. One of these serves to select the correct locomotive decoder (Märklin or Elektor Electronics), while the other provides the additional function required by the Märklin decoders.

Each controller enables one of the possible 81 addresses to be set by a diode matrix. To keep the wiring to a minimum, address setting is carried out on the main board by mounting a number of diodes. A spartan but inexpensive solution if you do not use more than 16 locomotives. The setting may be carried out in a somewhat more flexible manner with the aid of DIL switches or jump leads, or in a very convenient manner by means of thumb-wheel switches. The latter is possible because the locomotive addresses are written in BCD format. Yet another possibility is a hybrid system in which thumb-wheel switches are used for a couple of controllers and diodes for the others. How you implement the addresses depends therefore on the money you are prepared to spend on it.

An even more de luxe solution is still being contemplated. This entails setting the addresses fully electronically and includes a two-digit seven-segment display for each locomotive.



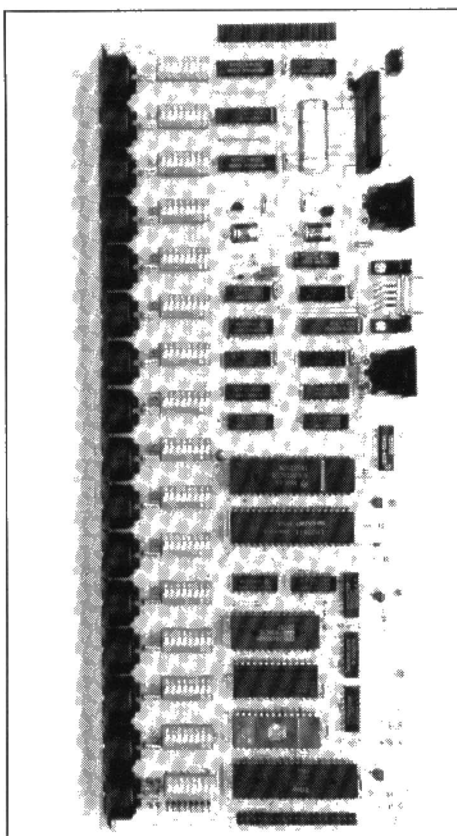


Fig. 32. Evolution of the main board of the Elektor Electronics Digital Train System—at the top the very first prototype and underneath it the final design.

Although only sixteen controllers can be connected to the main board, it is possible to operate 81 locomotives simultaneously. Just as in the Märklin system, any controller may set the speed of a given locomotive and then be switched to another locomotive. The first locomotive will, of course, continue to move at the set speed.

It is also possible to control locomotives that are not connected to a controller via the RS232 interface.

A third possibility is accessing the controllers by the host computer via the RS232 interface. The host computer processes the data and retransmits them to the control system. In this way, several locomotives may be operated via one controller (for instance, when two or more are pulling heavy goods trains).

It is also possible to simulate the inertia of trains or to restrict trains to a certain maximum speed.

The main board contains a dead-man's switch, which, when pressed,instantly stops the data transmission to the rails so that all rolling stock comes to a standstill.

Booster

The Märklin Central Unit contains an integral power amplifier (booster) that can deliver up to 3 A. Our system provides rather more power, since each locomotive draws 0.5–1.0 A.

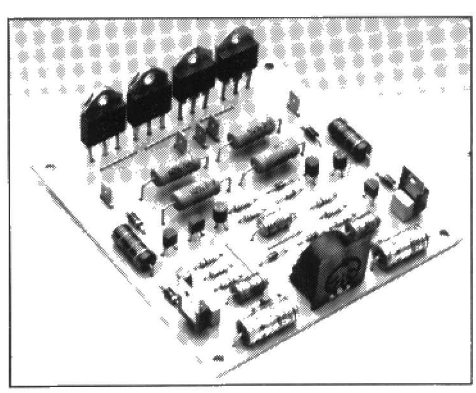


Fig. 33. A separate booster stage supplies ample power to the track.

Our booster is designed as a separate unit and provides a voltage-stabilized output. It may be used with the Märklin system. Stabilization of the output is obtained

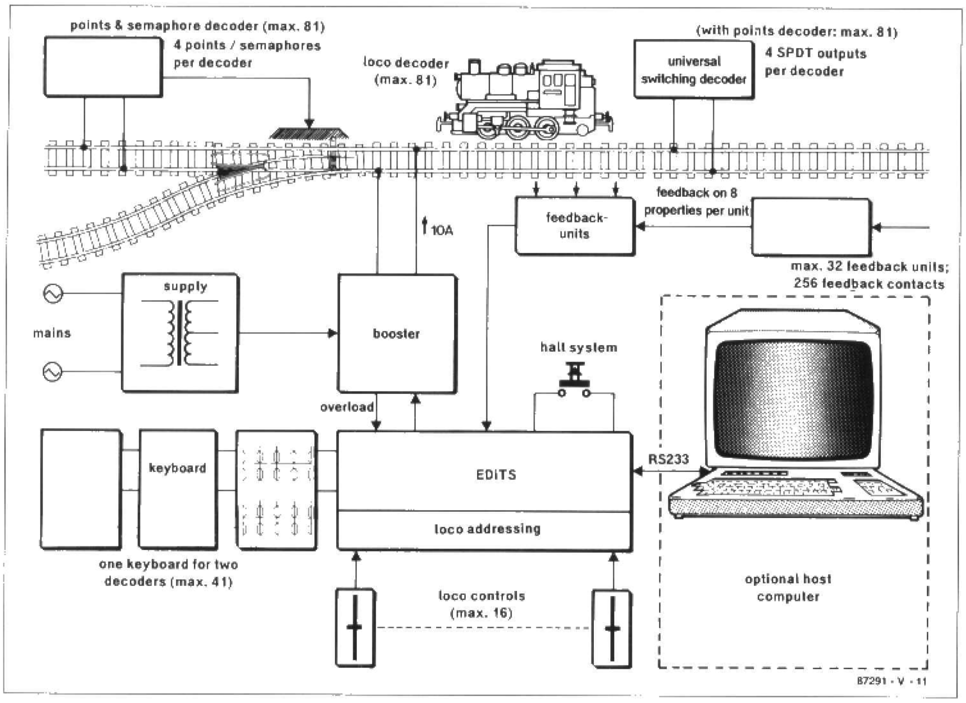


Fig. 34. Block schematic of the complete Elektor Electronics Digital Train System.

EEDTS	Märklin	Remarks
Turnout (points) decoder (Feb 1989)	Turnout (points) decoder (K83)	Suitable for four turnouts (points) or signals or for eight sorting or marshalling sidings
Locomotive decoder for d.c. and a.c. systems (Mar/Apr 1989)	Locomotive decoder. C80 for a.c. C81 for d.c	
Two-rail adapter (Mar/Apr 1989)		Märklin has announced a separate locomotive decoder for two-rail systems
Universal signals and switching decoder (May 1989)	Switching decoder K84	EEDTS offers various output configurations: darlingtons or relays
Main PCB (June 1989)	Central unit + Control 80 + interface	EEDTS PCB has integral RS232 and 16 locomotive controllers
Booster (10 A) (July 1989)	Booster (2.5 A)	EEDTS booster is stabilized
Keyboard(s) (Sept 1989)	Keyboard(s)	Not compatible
Monitoring unit for 8 sensors (Oct 1989)	Monitoring unit for 16 sensors	Not compatible
	Memory	

Table. 4. Brief comparison of the Elektor Electronics Digital Train System and the Märklin system.

by configuring the output stage as emitter follower. This has the advantage that the output voltage is independent of the input voltage and will therefore remain stable, even under heavy load conditions.

On the track this is manifested by truly constant lights and well-defined maximum speeds of the locomotives.

The booster output is proof against short-circuits to prevent fires in case of a derailment or other accident. When the output current exceeds 8–10 A, the protection circuits are automatically switched on. If the short-circuit is of short duration, the protection circuits are switched off again automatically. When a short-circuit persists, the dead-man's switch is operated. This effectively prevents overheating of the booster and a possible fire hazard on the track. If you have a very large track, you can, of course, use a number of boosters.

Keyboards

In principle, it is possible to use our system for the independent control of locomotives only, while signals and turnouts (points) are operated in a conventional manner. The system has, however, been designed to control not only locomotives, but also signals, turnouts (points), sorting sidings and any other switching functions you may have. You do, of course, need a number of relevant decoders as described in previous articles, although Märklin units may also be used. For every two decoders one small keyboard containing 16 keys is required. Therefore, if your system has the maximum 81 locomotives, you need 41 keyboards.

The 16 keys on the keyboard serve to operate the eight signals or turnouts (points) that are connected to the decoders associated with the keyboard. The keyboards provide a visual answer-back signal in the shape of LEDs to keep you informed of the actual position of signals and turnouts (points).

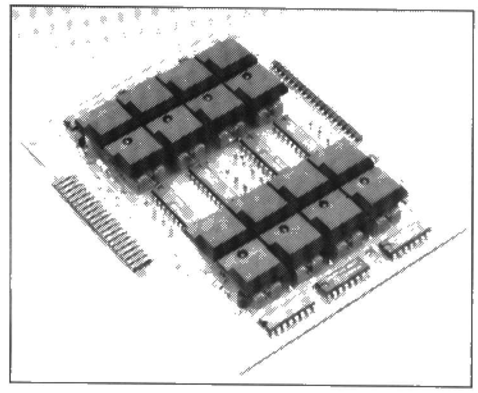


Fig. 35. The Elektor Electronics Digital Train System may be expanded with keyboards: how many depends on the number of signals and turnouts (points) the track contains.

Each keyboard has a 20-pole connector at each of its sides to enable a number of them to be interlinked without any wiring. The extreme right-hand keyboard is plugged into the relevant connector on the main PCB.

You are not tied to the layout of a keyboard: it is perfectly possible to incorporate the switches in an operating panel.

The advantage of controlling signals and turnouts by means of decoders and not in a conventional manner was already discussed in the first article in this series. The advantage of switching them via the present system is that they may be managed centrally. The actual position of the various signals and turnouts (points) may be ascertained via the serial RS232 connection. Moreover, switching instructions may be given via this route, that is, by-passing the keyboards. These features are indispensable if you want to operate a software-controlled protection system or timetable

via the host computer. Thanks to the system, your computer does not need a (parallel) output for every signal and turnout (poins): it is all done via that one RS232 connection.

Monitoring units

Apart from keyboards and controllers, the system can also incorporate a number of monitoring units. These units make it possible, for instance, to locate a locomotive anywhere on the track. Each monitoring unit has provision to accept up to to eight sensors. A number of units may be connected in parallel via a five-way cable. The units may be located along the track in a decentralized manner. Note, however, that our monitoring units are not compatible with those of Märklin.

The signals from the monitoring units may be written into the computer via the system on command. Since each of the units has its own local memory, even very short switching pulses will not be missed. The writing and processing of monitoring signals can take place only via the RS232 interface. The host computer connected to the interface can be programmed in a manner to cause certain monitoring signals to result in predetermined switching operations.

RS232 interface

The RS232 interface fitted as standard on the main board offers a multitude of possibilities. It is, unfortunately, not possible to give a detailed description of these, because at the time of writing the software for the control program (which must, of course also interpret the RS232 commands) was still in development. At present we can therefore only say that we intend to give the RS232 interface the following facilities:

- giving control commands to locomotives (speed; forward; reverse; additional function of Märklin locomotive decoder);
- obtaining information on the position of connected locomotive controllers and the associated locomotive addresses;
- releasing or otherwise of certain locomotive controllers;
- giving switching instructions for the control of turnouts (points), signals and sorting sidings;
- actuating switches operated via the decoders;
- releasing or otherwise of keyboards: in fully automatic operation the keyboards, like the controllers, may be locked by the program;
- obtaining information on the position of signals and turnouts (points);
- obtaining information from the monitoring units;
- reset: all locomotives at standstill although power to the track is maintained;
- emergency stop: power is removed from track (this may be an optional facility);
- setting of the baud rate of the RS232 interface;
- setting of duration of actuation of turnouts (points) (when these are controlled via the keyboards, they are actuated as long as the key is depressed; when they are controlled via the RS232 interface, a fixed actuation period must be set to prevent their coils burning out);
- loading of application programs from the host computer—real enthusiasts can write their own application programs in Z80 machine language and load them in the RAM of the system; the programs can be started via separate instructions.

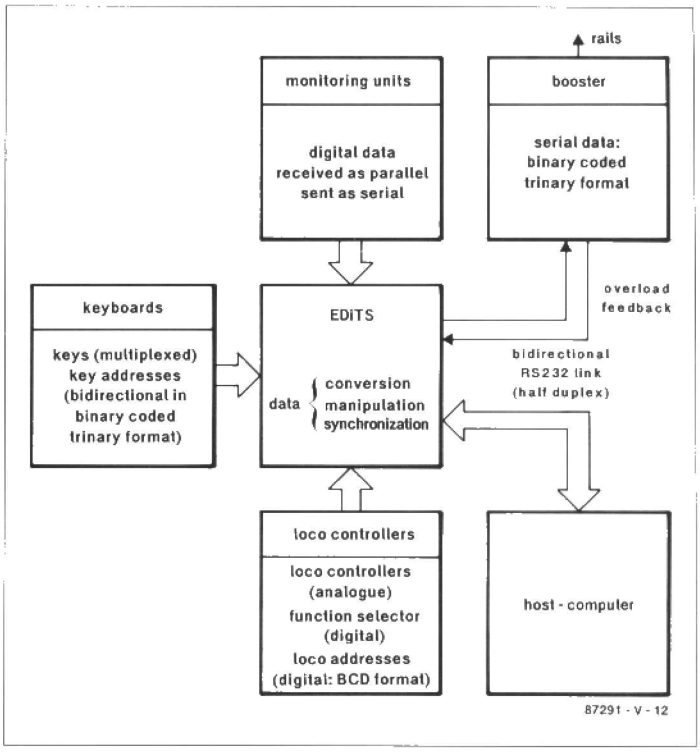


Fig. 36. Signal flow diagram showing the various data formats used.

This is all we can say about the RS232 interface, but the subject will be reverted to later in the series.

Response time

The data flow in the system is primarily serial. Because of the uncertain contact between locomotives and rails, control instructions to the locomotives are transmitted constantly, but only to those locomotives that are in operation. Unused controllers are recognized by the system and ignored. The associated locomotive addresses are defined as out of service, except if they are in use via the RS232 interface. Only the data of actually operating locomotives are sent out via the rails.

The result of this procedure is that the response time of the system depends to a large extent on how many locomotives are in operation at any one time. When all 81 locomotives are in use, the response time will be of the order of 0.8–1 second.

Switching instructions for decoders along the track are always transmitted, however, in between the locomotive commands, ensuring that they have a response time not exceeding about 10 ms. Each switching instruction is transmitted only twice, but since the decoders are connected to the track permanently the information transmission is reliable.

Control program

The Elektor Electronics Digital Train System is based on a Z80 microprocessor. The control program is being developed in assembler and will become available through our Readers' Services in due course.

Although it is not necessary for the user to know exactly what the program does, it is, of course, interesting to know at least the basics of it, if only to get a better insight of how the system works.

The main task of the program is communication. Information is obtained from, and transmitted to, a number of locations and the system must be able to co-ordinate this flow of information and transform it into switching instructions that are passed to the locomotives, signals and turnouts (points) via the rails.

Control instructions are transmitted via the locomotive controllers, keyboards, monitoring units and the RS232 interface. These instructions are of different format: they may be analogue (locomotive controllers, for instance) or digital. The digital data may be parallel (locomotive addresses in BCD format, keyboard addresses in binary coded trinary format) or serial. The serial data representing voltage levels, baud rates and protocols are all different from one another.

The first task of the system is, therefore, the translating and conversion of these different data formats.

Another aspect is that the data are asynchronous, that is, they are obtained or transmitted at different times. However, the serial output can handle only one instruction at a time. Synchronization and allocating priorities to control and switching instructions are, therefore, another vital task of the control program.

Program and data structure

The control program consists, strictly speaking, of two different modules that are vying for attention on an interrupt basis. The Elektor Electronics system is, therefore, a multi-tasking system. The main task

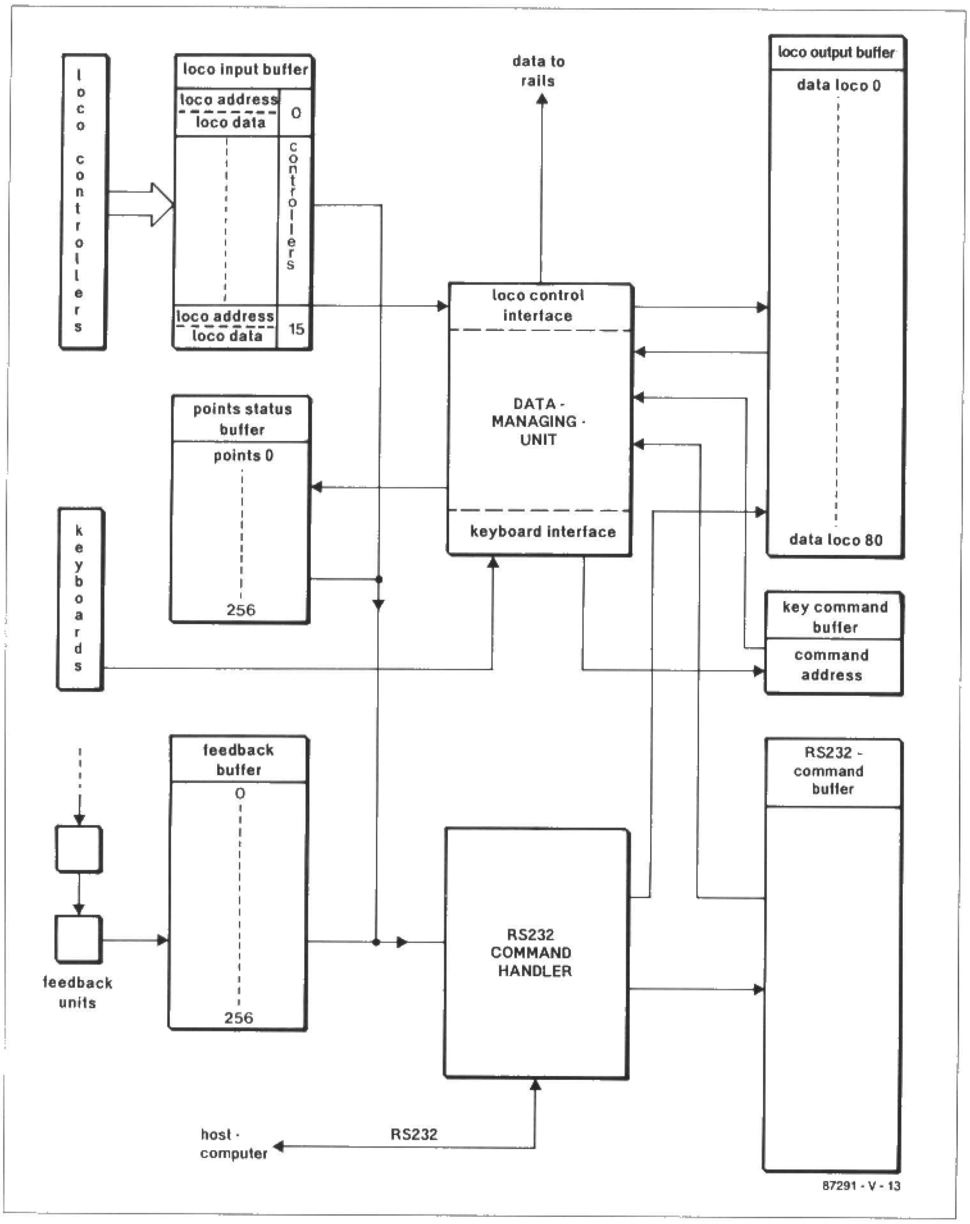


Fig. 37. Internal data structure of the control program. Two separate modules handle the data processing: the data managing unit and the RS232 command handler.

is performed by the data managing unit. This part of the program carries out the loading and processing of the controller and keyboard data and ensures that these are transmitted via the rails in good time.

The second module, the RS232 command handler, interrupts the data managing unit as soon as RS232 instructions appear on the line. It interprets the received commands and ensures that the system undertakes correct action for each instruction.

All incoming data are collected, processed if necessary and put into sequence before they are retransmitted via the rails. In this respect, the system resembles a post office. In the RAM a number of buffers have been reserved for sorting data. Other buffers merely collect incoming data and yet others hold data ready for transmission.

The internal data structure is shown in Fig. 37, which also shows the position of the two modules and to which buffers they have access.

As already mentioned, the data managing unit performs the routine tasks. It loads the controller and keyboard data in the locomotive output buffer and key command buffer respectively. It also carries out any required data conversions and, for instance, adaptation of Elektor Electronics or Märklin locomotive data formats. Moreover, it retransmits data via the rails, including any stored in the RS232 command buffer.

The RS232 command handler takes care of the communication with an external host computer. On the one hand it contains module routines with which the RS232 interface is realized (the system does not use a special RS232 chip) and on the other hand it contains a decoding routine to decipher incoming commands.

Incoming locomotive control instructions are placed direct into the locomotive output buffer. This means, in effect, that the data managing unit loses control of the associated locomotive address until this is released again by the RS232 interface.

Switching instructions are loaded into a separate buffer. Yet other buffers are used by the RS232 command handler for monitoring or ascertaining the position of signals or turnouts (points).

The RS232 command handler has access also to the locomotive input buffer for monitoring and ascertaining the position of controllers and set locomotive addresses.

The monitoring (feedback) buffer is accessible only via the RS232 interface. If it is necessary that a given monitoring signal requires a certain action, this has to be programmed via the host computer. This means that if you do not intend to connect the system to a computer, there is no sense in using monitoring units.

Cost aspects

Before you start work on a complete and far-reaching digitization of your model railway, you would, no doubt, appreciate what sort of outlay you may expect.

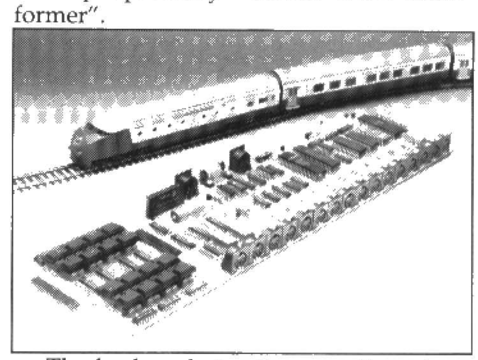
A first glance at the main PCB may make you hesitate even to begin thinking about starting, but, although it looks complex (and therefore expensive?), we believe that the proposed system is very inexpensive for what it offers.

The main expenditure will almost certainly be the main PCB (for price, see Readers' Services section), which is double-sided and through-plated (not easy to make yourself or even to have made). But almost all the components used on this board are fairly, or very, cheap.

Also bear in mind that the locomotive addresses may be set with the aid of cheap diodes. More sophisticated means (DIL switches, for instance) are entirely your own choice.

Another point worth remembering is that connecting the locomotive controllers via DIN connectors is, strictly speaking, an unnecessary luxury.

The main expense in the booster lies in the mains transformer and the heat sink, but even for a non-digital track you need at least one transformer. Here, it is worth bearing in mind that it is invariably much cheaper to buy an appropriate transformer from an electronics retailer than to insist on a proprietary "model train transformer".



The keyboard circuits, as well as those for the monitoring units, have been kept as simple and inexpensive as feasible, particularly since we realize that most users will want (or need) a number of these units.

OPTICAL SHAFT ENCODER FROM SHARP

T. Wigmore
source: Sharp Electronics

An optical shaft encoder is an electronic device used in applications where the position or speed of a rotating shaft needs to be known for processing and control purposes. Practical uses of this interesting device may be found in tool control, robotics, positional control systems, and many more. This article introduces the Type GPIR04 positional sensor from Sharp Electronics, and discusses some of its practical implementations.

Measurement and control of shaft rotation can be implemented in a number of ways. The classical tachometer is fine as a revolution counter, but gives unsatisfactory results at low shaft speeds. The instrument is also less suitable for shaft position measurement because the output signal is a function of the rotational speed. For position measurement, the analogue signal supplied by the tachometer requires to be integrated. Apart from the difficulties that may be expected to arise from the non-ideal behaviour of analogue integrators, accumulation of errors must be taken into account. An integrator works on the principle of addition, and is normally incapable of discriminating between the instrumentation signal and the error signal.

The above considerations have lead engineers to resort to the potentiometer as a more appropriate means of achieving positional feedback. The resistance of the potentiometer, whose spindle is secured to the shaft, changes as a function of shaft position. This solution has its drawbacks, too: in many cases, it is difficult to gear the full travel of the shaft to the limited turning range of the potentiometer spindle. Another serious point to be taken into consideration is the wear of the potentiometer.

For shaft position and speed measurements, the optical encoder has long been known to have none of the disadvantages associated with the tachometer and the mechanically coupled potentiometer. Until recently, however, the price of incremental encoders was prohibitive for many electronics enthusiasts. Most encoder types used in industrial applications cost £20 to £40.

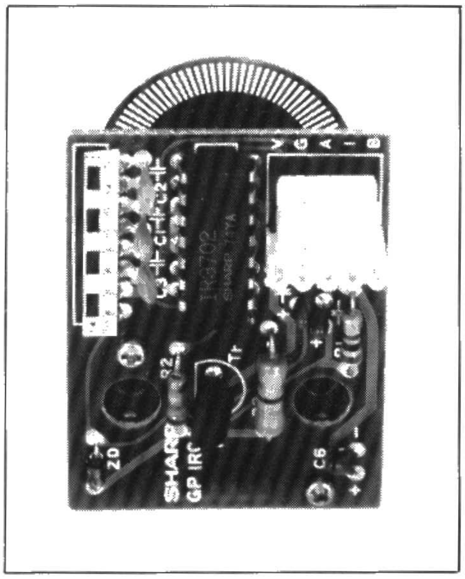


Fig. 1. The optical shaft encoder complete with associated interface board.

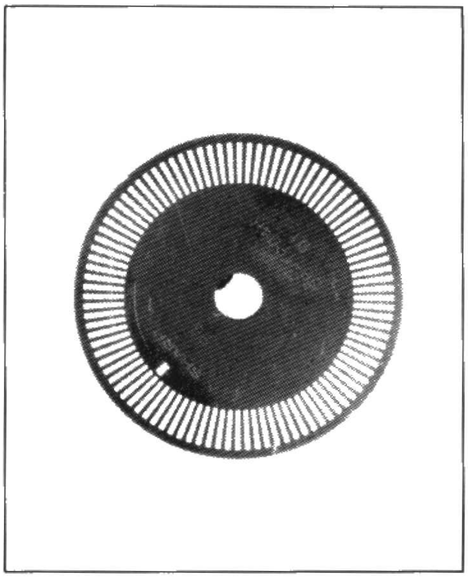


Fig. 2. Pulse disc with 96 slots and a shaft hole of 4 mm diameter. Note the additional hole for the synchronisation pulses.

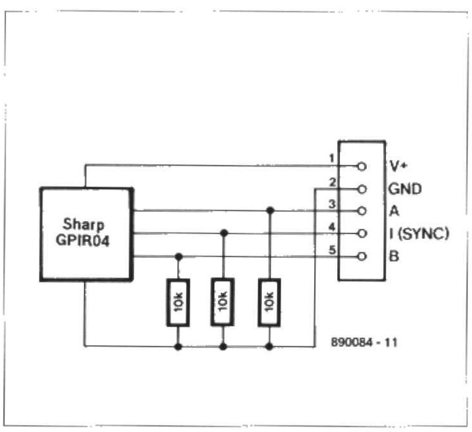


Fig. 3. Basic electrical connection.

Affordable

Sharp Electronics have recently launched a versatile shaft encoder whose cost is lower than any comparable device used earlier in industrial applications. The new GPIR04 is a shaft encoder of the *incremental* type.

An incremental encoder produces output pulses as a function of shaft rotation. These pulses are generated by a slotted optocoupler to eliminate the problem of wear and mechanical loading of the shaft. A circuit based on this principle was published in Ref 1. The home-made construction described in this article has the disadvantage of suffering from limited resolution, because here this depends largely on the accuracy used to produce the pulse disc.

Professional pulse encoder discs are manufactured by etching and laser trimming techniques and naturally offer much greater accuracy than home-made versions.

The pulse disc produced by Sharp has a very fine structure which makes the device look like the head of an electric shaver (Fig. 2). The sample encoder received from Greenweld Electronic Components had 96 slots, but types with 100 slots are also manufactured.

As shown in Fig. 3, the encoder has three outputs. The pulse frequency and phase relation of two of these indicate rotation speed and direction of travel of the shaft respectively. The other output supplies a pulse at each complete shaft revolution, and is used for synchronization

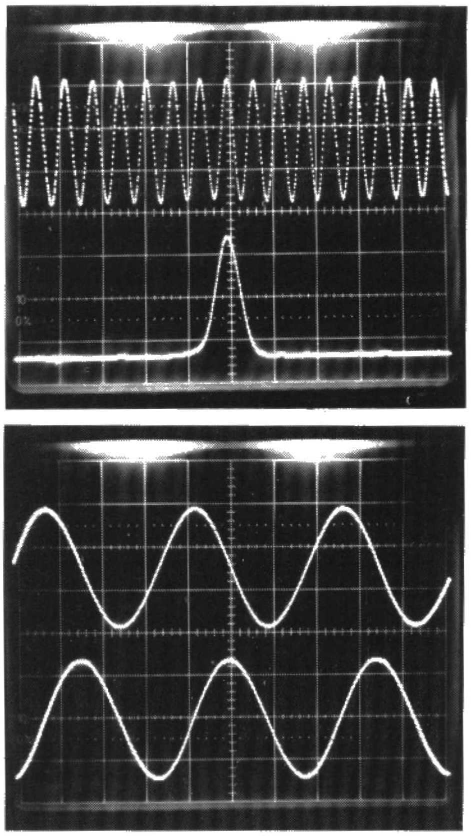


Fig. 4. **Analogue output signals supplied by the shaft encoder.**

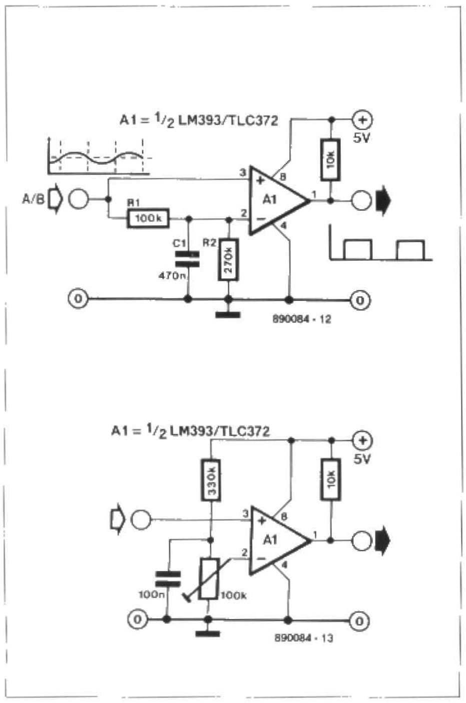


Fig. 5. **Two suggestions for an interface circuit between the encoder and a digital circuit such as a counter.**

purposes. The maximum achievable pulse frequency is stated as 20 kHz, which corresponds to a shaft revolution rate of 12,000 per minute.

Although the encoder works from a 5 V supply, its output signals require conversion before they can be applied to a digital circuit. A simple interface such as the one drawn in Fig. 5a provides the required function. The comparator circuit used has the advantage of comparing the signal supplied by the encoder with its own average output signal rather than with any predefined threshold level. This results in automatic compensation of DC level shifts of the encoder output signal caused by, for instance, ambient light changes. The disadvantage of the circuit of Fig. 5a is that it fails to work correctly at very low input frequencies (= shaft speeds). In that case, the alternative shown in Fig. 5b gives better results.

As shown in Fig. 6, the position of the pulse disc can be fine-tuned with respect to slotted optocouplers. This precise adjustment is required to optimize the amplitude of the output signal, and to establish an accurately defined phase relationship. The latter parameter is of great significance for increasing the resolution, which will be gone into a little further.

The direction of rotation is fairly simple to establish on the basis of the phase relation between signals A and B, and with the aid of a J-K bistable (see Fig. 7.) Gates N_1 and N_2 in this circuit provide an OR function for signals A and

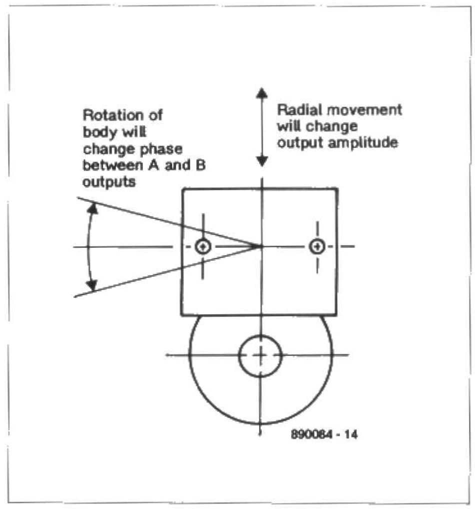


Fig. 6. **Adjustment of the encoder body for optimization of the phase difference and the output amplitude.**

B, and in addition ensure the required, short delay of the clock signal for the bistable. The J and K inputs are thus given enough time to stabilize. Gates N_3 and N_4

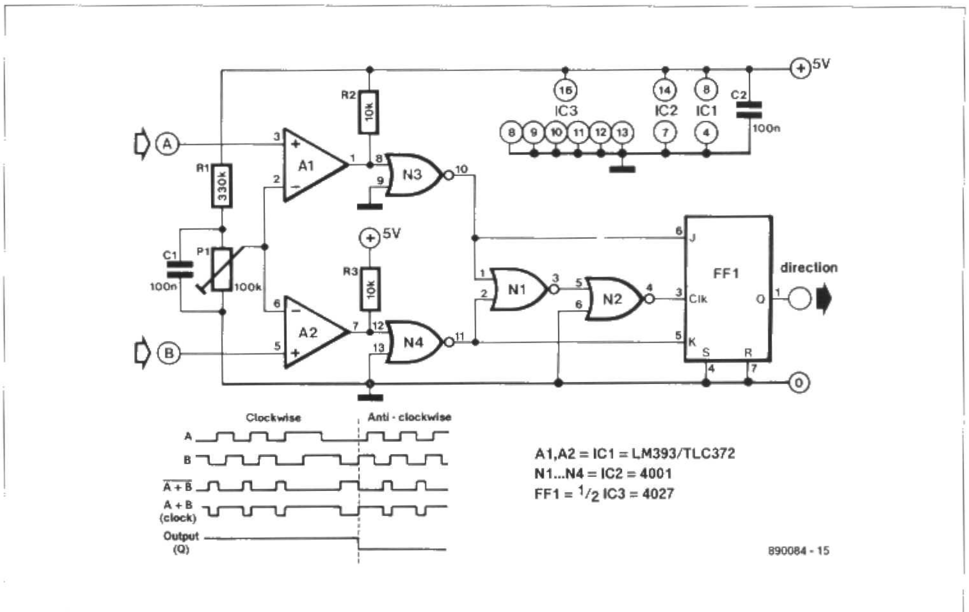


Fig. 7. **Decoder circuit for deducing the direction of rotation.**

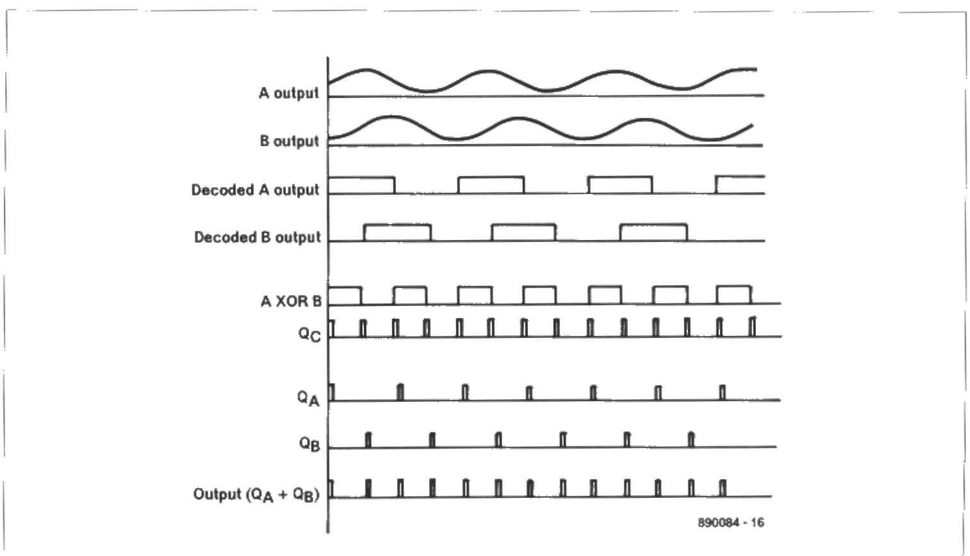


Fig. 8. **The number of pulses per shaft revolution may be quadrupled by combining signals A and B, and detecting the pulse edges. The 5th and 6th timing diagram from the top refer to Fig. 9, and the lower three to Fig. 10.**

Tabel 1

Absolute maximum ratings

Supply voltage	6 V dc
Current consumption	50 mA
Operating temperature range	0 to +70°C
Storage temperature range	−20°C to +80°C

Electrical characteristics at 25°C, V_{OUT} = +5 V, R_L = 10 kΩ

Parameter	Min.	Typ.	Max.	Units
Pulses per output (per revolution)		96		
Phase shift between A and B outputs	70	90	110	°
Output signal A & B (see Figure 1)	1,0			V
Sync. pulse output (see Figure 2)				
Low level			0,5	V
High level	1,0			V
Operating frequency			20	kHz
Equivalent disc speed			12.000	rpm

Mechanical characteristics at 25°C

Parameter	Min.	Type	Max.	Units
Allowable run out on disc			0,1	mm
Allowable concentricity disc to shaft			0,1	mm
Disc outer diameter	26,98	27,00	27,02	mm
Disc hole diameter		4,00		mm
Disc thickness		0,1		mm
Allowable angular misalignment disc to encoder datum face			5	°
Disc material	Stainless steel			

Table 1. Electrical and mechanical specification of the GPIR04 optical shaft encoder.

improve the edge steepness of the output signal supplied by A1 and A2. In a comparator of this type, with open-collector or open-drain outputs, the 0-to-1 change time is fairly long and dependent on the value of the pull-up resistors used. If a comparator (or opamp) with a symmetrical output is used, the pull-up resistors and N3-N4 are omitted. A practical example based on the TLC272 is given in Fig. 9.

Increasing the resolution

The 96 pulses supplied by the GPIR04 determine the smallest measurable shaft rotation. The resolution, however, is greater than 1/96th revolution, because pulse edges rather than pulses can be counted. Establishing the difference between signals A and B with the aid of a XOR function, for instance, results in 192 symmetrical pulses per revolution (see Fig. 8). This number may be doubled to 384 pulses per revolution simply by processing the pulse edges rather than the pulses themselves. The same 384 pulses per revolution may also be obtained by adding up the short pulses generated on the edges of the decoded A and B signals.

Reference:

1. Revolution counter. *Elektor Electronics* September 1981, pp 9-02–9-06

The sample optical shaft encoder Type GPIR04 and its datasheets were kindly furnished by Greenweld Electronic Components • 443 Millbrook Road • Southampton SO1 0HX. Telephone: (0703) 772501/783740.

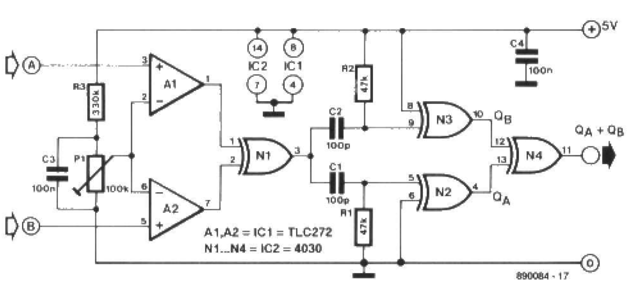


Fig. 9. Basic design of of a pulse multiplier. The signal at point A (192 pulses per shaft revolution) is still symmetrical.

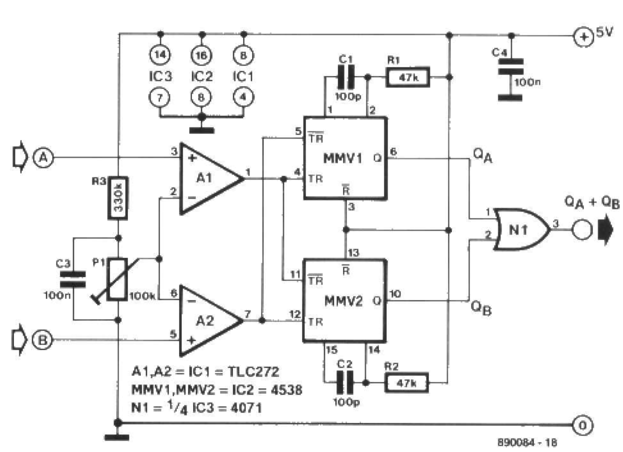


Fig. 10. An alternative to the circuit of Fig. 9.

RAM EXTENSION FOR BBC-B COMPUTER

P. J. O'Shea

The original BBC microcomputer has one serious shortcoming: lack of user RAM. In the highest screen resolution mode, on a disk system, there are not quite 6 Kbytes available for BASIC programs. The RAM extension board described in this article raises this figure to over 25 Kbytes, available to any language, including word processors, independent of screen mode and without reducing processing speed. Driver software is available to support the extension, which can be used as two banks of sideways RAM, into which sideways ROM programs can be loaded without using up a ROM socket inside the machine.

To the user of shadow RAM it appears as if the extension memory is the screen memory and external to the microprocessor memory field. Hence the whole of what used to be the screen memory is freed for use.

In reality, the shadow memory appears in the memory map as shown in Fig. 1. The shadow block is sideways to a block of the main memory in the range 2000H to 8000H, i.e., either the main memory or the shadow memory, depending on which of the two is activated, can be addressed by the microprocessor in this range. The screen memory is always in the main block, which is physically located on the main circuit board, since it is closely linked to the screen display circuits that continuously read (and thus refresh) it.

To make the shadow RAM available for BASIC requires a control ROM which performs the following function. Ordinarily, the shadow RAM block is selected, displacing the screen memory, but the screen memory is recalled briefly when required: this occurs mainly during the OSWRCH (*Operating System Write Character*) VDU routine. All graphics control commands available in BASIC are executed simply by passing bytes to OSWRCH. The OSWRCH vector is intercepted — in fact, the extended vector must be used to make OSWRCH calls pass into the controller (paged) ROM. To service a call, the screen memory is selected, after which the normal OSWRCH service in the operating system ROM is invoked. Finally, the shadow RAM is re-selected.

The hardware

The RAM board plugs into the microprocessor socket in the BBC micro, from which it draws its power. The original microprocessor is moved to a socket on the RAM board. Four other connections must be made to the BBC micro circuit board.

Whether addressed as shadow or sideways memory, the RAM is required to take priority over memory already existing in the system — the main RAM, if addressed as shadow RAM, or an echo of a sideways ROM caused by incomplete ROM-select decoding if addressed as sideways RAM. The effect of the existing memory is removed by isolating the data bus with a three-state buffer when access to the extension RAM is required. In addition, a 'read' is signalled on the system read/write line so that the main RAM can not be corrupted.

Addressing as sideways RAM

Inside the BBC micro, a 4-bit latch, IC_{76}, is used to select one of 16 possible sideways blocks (in the memory range 8000H to BFFFH). The four internal ROM sockets are enabled by a decoder connected to the two low bits. The sockets from left to right are activated by the two low bits indicating 0 to 3 respectively, and are usually considered by the operating system as block numbers 12 to 15. One of the outputs from this decoder (taken from the CHIP ENABLE pin of one of the four ROM sockets) and the two high bits of the ROM-select latch (pins 12 and 11 of IC_{76}) account for three of the four wired connections to the BBC micro's main circuit board.

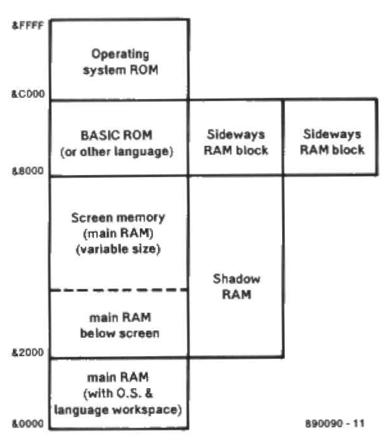


Fig. 1. Memory map of the BBC microcomputer with sideways and shadow RAM blocks.

The sideways RAM is activated by the ROM Chip Enable signal if the highest (fourth) bit of the ROM-select latch is low. The third bit then discriminates between the two 16 Kbyte sideways blocks. Hence if the connection is made between the Chip Enable signal of the ROM socket at the extreme left, the sideways blocks are numbers 0 and 4, while the socket to the right of this would give block numbers 1 and 5, etc.

Addressing as shadow RAM

The upper 8 Kbytes of the lower sideways RAM block and the whole of the upper block double as the shadow RAM. The extension RAM is always addressable as sideways RAM, but is addressable as shadow RAM (in the range 2000H to 7FFFH) only when selected. Selection and de-selection are achieved by performing a read of the ROM-select latch addressed at FE30H to FE3FH. The shadow RAM is selected by reading from any odd-numbered address in this range, and deselected by an even address, i,e., the lowest bit of the address bus is latched into the Shadow Enable flip-flop.

The fourth connection to the BBC main board is to an otherwise unused decoder output, IC_{26} pin 7, which goes low to indicate a read in the above range. Note that

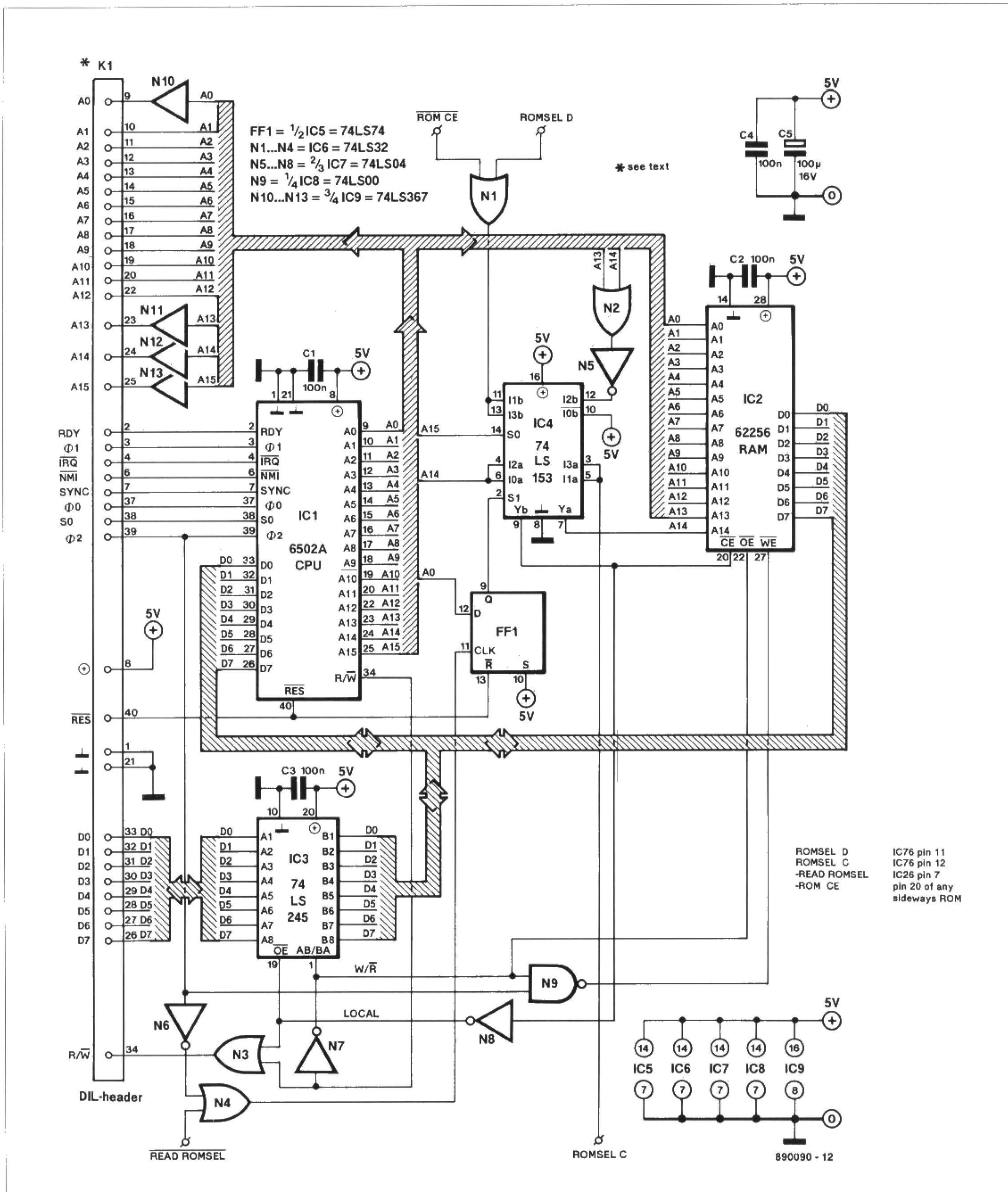


Fig. 2. Circuit diagram of the RAM extension card. The microprocessor is taken from the computer's main board.

the ROM-select latch is not affected by the read action.

Control program

A listing is available that enables the control program to be reproduced with the aid of the BASIC assembler. Level 2 BASIC is required since the assembler incorporated in it provides the facility to produce object code at one location that is intended, however, to run at another. Clearly, the assembler can not assemble directly to sideways RAM, since the BASIC interpreter displaces all other sideways blocks. This is why the code is assembled to an area of main RAM, and then copied to sideways memory. Here it is secure, since it fits into the lower 8 Kbytes of the lower sideways block (which is not used as shadow memory). Blowing the control program into EPROM is therefore not absolutely necessary, but is obviously convenient because it obviates the need to load it from tape or disk before using the shadow RAM. When used, the EPROM must be a type with an access time of

```
 10 REM*** SHADOW RAM CONTROL ***
 20 REM* requires OS 1.2
 30 :
 40 REM*  to save object code (for 2764 EPROM) use:
 50 REM*   *RSAVE filename <RNUM>
 60 REM*  where RNUM is as follows:
 70 RNUM=0: REM* Set number of lower sideways RAM block, 0-3
 80 :
 90 REM* commands:
100 DATA ROMS,RLOAD,RSAVE,SHADOW,*
110 :
120 REM* 'help' text: (÷=CR,LF  °=<8 spaces>)
130 DATA (C=check RAM) list sideways ROMs
140 DATA <filename> <block number 0-F>÷°load to sideways RAM
150 DATA <filename> <block number 0-F>÷°save from sideways memory
160 DATA (0=shadow only÷°  1=shadow if MODE 128-135÷°  2=shadow off)
170 :
180 PROCREADCOMS: REM* Read above data
190 DIM R% &700
200 NAME$="Shadow RAM":RTYPE=&82: REM* ROM name & type code
210 :
220 ABCODE=&AE0: REM* a few bytes required for some OSWORD calls
230             REM* in a cassette system, use ABCODE=&D01
240 CODE=&E00: REM* main ROM workspace
250 RVEC=&F6:MVEC=&F8: REM* misc. vectors
260 :
270 SHON=&39F:REM* bit7=Shadow mode,bit6=Inside OSWORD,bits1&0=SHADOW option
280 LCHAR=&3A0:REM* last character to OSWRCH
290 :
300 REM** O.S. Variables:
310 PARBL=&2EE: REM* Parameter block for *LOAD/*SAVE
320 CURLANG=&28C:REM* current language
330 BASROM=&24B:REM* BASIC ROM number
340 NUMCH=&26A:REM* length of VDU queue
350 BKTYPE=&28D:REM* zero=last BREAK was soft
360 SMODE=&355:REM* screen mode
370 :
380 OSASCI=&FFE3:OSNEWL=&FFE7: REM* O.S. routines
390 OSBYTE=&FFF4
400 OSRDRM=&FFB9:OSFIND=&FFCE
410 OSFILE=&FFDD
420 :
430 EXVEC=&D9F:REM* extended vector storage
440 DEFFNEXV(V%)=EXVEC+(V%-&200)DIV2*3: REM* returns position of
450       REM* extended vector given position of absolute vector
460 :
```

Fig. 3. Extract from the control program developed by the author.

250 ns or faster.

Whether in RAM or EPROM, the control program is best placed in a higher priority (i.e., higher-numbered) sideways ROM block than the filing system ROM, so that the shadow memory remains active after a break condition. Note that the priority of the ROM sockets on the BBC main board increases to the right of the row. The RAM driver program allows the filing system to initialize as usual, whereas the filing system will not allow any ROMs below it to initialize on break.

Unless a ROM extension board is in use, it will not be possible to place the filing system ROM in a lower priority block than the lowest RAM block. With the selection logic described above, the lowest RAM block number must be in the range 0 to 3. This provides another reason why the use of EPROM to hold the RAM driver program is more convenient.

Other software details

There are two vectors other than OSWRCH that must be intercepted to ensure that the screen memory is present when required: OSBYTE (call number 135 reads a character from the screen) and OSWORD. OSWORD call 9 reads a pixel — this call is made by POINT in BASIC — and a call number zero inputs a complete line. The screen memory is required for the latter not because the characters typed appear on the screen (they are passed to OSWRCH), but since the use of the COPY key causes the reading of characters without calling OSBYTE 135.

It should be noted that some OSWORD calls cause disk access, such as saving or loading to memory — the screen memory must, therefore, not be recalled to service these. Note also that on some odd occasions a program in sideways ROM may use an OSWORD parameter block in that same ROM, although BASIC avoids doing this. If this is the case, one answer is to establish a routine in absolute memory, outside of the paged ROM system, which restores the calling ROM for the duration of the OSWORD call.

There is another reason for trapping OSBYTE calls: all languages use OSBYTE calls &84 and &85 to determine the upper limit of free memory, known as HIMEM in BASIC. This normally has a value between 3000H (Mode 0) and 7C00H (Mode 7). In a shadow mode, the control program returns 8000H in response to these calls. This means that the extension RAM is made available to any language.

Data is shifted between the main RAM and shadow RAM on changing the Mode type to prevent loss of the BASIC program.

In the listing of the RAM driver software, the code to implement shadow RAM appears between lines 640 and 2070. The remainder of the code implements several utilities that are described below.

Control commands

To maintain compatibility with existing software, the shadow memory is not automatically enabled at power-on. Modes 0 to 7 are designated as non-shadow modes, in which the machine keeps the usual internal configuration. Shadow memory is enabled by selecting Modes 128 to 135 which give shadow versions of Modes 0 to 7 respectively. In all shadow modes, the BASIC variable HIMEM returns &8000.

The mode type, i.e., shadow or non-shadow, is preserved by a soft break, and reset to non-shadow by a hard break (CTRL + BREAK).

The utilities implemented in the control ROM are as follows. Command *SHADOW results in only shadow modes being selected so that selection of Modes 0–7 actually gives 128–135. Command *SHADOW 1 returns normal operation. *SHADOW 2 ensures that only non-shadow modes are selected — this can be used to protect the RAM if it is to be used as sideways memory. The *SHADOW option is preserved even through a hard break.

Other utilities include *RLOAD which loads a file to a sideways RAM block, and *RSAVE which saves from sideways RAM to ROM. Note that these commands corrupt user memory in the manner of some DFS commands, and that *RSAVE always saves the entire 16 Kbyte block. Command *ROMS provides a list of all 16 sideways blocks, stating ROM or RAM and naming any program therein. An automatic RAM checking facility is included, and enabled by command *ROMS C.

All the commands are listed, and the syntax given, by *HELP SR.

In practice

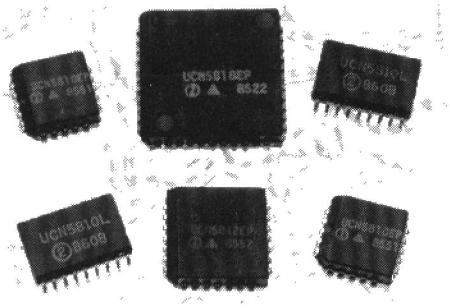
The prototype of the RAM extension was built on a Vero DIP board sized 2.8" by 4.0" (approx. 7×10 cm), with the microprocessor sited at the lower right. Connection to the BBC microprocessor socket was made by a discrete-wire DIL header rather than the more convenient IDC type. It was found necessary to first insert a turned-pin DIL socket, and then plug the header into this.

The board was held mechanically in a horizontal position by making the links from the corners in the header of 20 SWG (dia. 1 mm) copper wire with suitable sleeving. The microprocessor runs hot, so sufficient overhead space must be allowed for convection.

Note: the listing of the driver software developed by the author and referred to above is available free of charge by sending a self-addressed, stamped envelope to our London office (overseas readers must include 2 IRCs for the return postage).

Smart BiMOS source drivers

A new series of smart microprocessor-compatible BiMOS II serial input latched source drivers is available from Sprague.

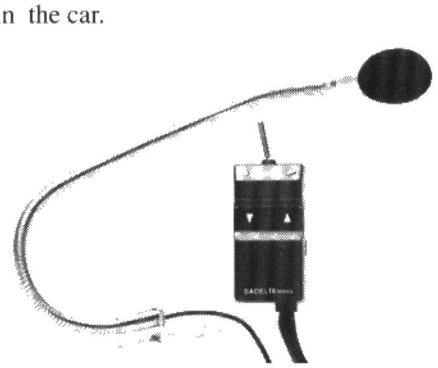


Designed primarily for use with vacuum fluorescent (VF) displays, the devices offer active DMOS pulldowns for high-speed ghost-free switching, reduced power requirements and a versatile choice of industry standard package size.

Sprague Semiconductors • Balfour House • Churchfield Road • Walton-on-Thames KT12 2TY.

Mobile Microphone

Nevada's Sadelta MM90 microphone is suitable for use with all existing bands of amateur transceiver and CB radio and has been designed to give hands-free operation in the car.

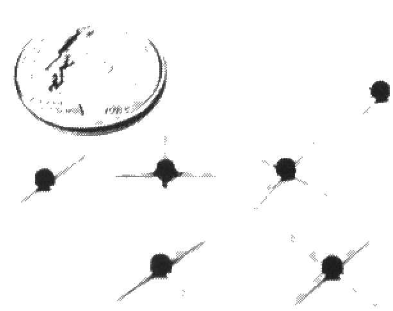


It uses an electret unidirectional insert for crisp audio quality and comes complete with a control box that allows up/down control of appropriate radios.

NEVADA • 189 London Road • North End • PORTSMOUTH PO2 9AE.

Low-cost GaAs FET

Avantek's low-noise GaAs FET for C- and Ku-band preamplifiers is now available from Wave Devices.



NEW PRODUCTS

Designated ATF-13284, the transistor is useful as a low-noise amplifier over the 2–16 GHz frequency range, and as an oscillator at up to 25 GHz. It offers 0.7 dB noise figure with 15 dB associated gain at 4 GHz, 1.6 dB with 8 dB gain at 12 GHz.

Wave Devices • Laser House • 132–140 Goswell Road • LONDON EC1V 7LE.

Signal Transmitter for Remote Computer Screens

A plug-in card for the IBM PS/2 range of PCs that allows monitors to be operated over long distances is available from Digivision. Type-coded VPSE 114, it is compatible with the VGA graphics standard and combines the sync pulses with the green video channel to generate a composite signal compatible with other systems.

Digivision Ltd • Parker Drive • LEICESTER LE4 0JP.

High-resolution ComputerDisplay Unit

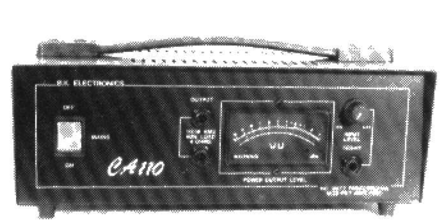
The 'Magus' high-resolution colour CRT monitor from Digivision works with most PCs and is particularly suited to applications in which such high-resolution is needed: in CAD and desktop publishing, for example.

The large screen (500 mm = 20 in. diagonal) offers a resolution of up to 1160 by 870 pixels to allow the clear display of fine detail. The horizontal frequency range is 15.5–37 kHz and that for the vertical is 48–90 kHz. The variation becomes important when the monitor is used with a multi-standard computer like the Archimedes, when the internal circuitry of the Magus will automatically adjust to display whatever format is sent by the computer.

Digivision Ltd • Parker Drive • LEICESTER LE4 0JP.

Cased Amplifiers from BK Electronics

In response to a large demand, BK Electronics have started to make their MF100 and MF200 modules available in a suitable enclosure.



The cases amplifiers will be known as the CA110 and CA210 slave amplifiers.

B.K. Electronics • Unit 5 • Comet Way • Southend-on-Sea SS2 6TR.

New Indoor TV Aerial

Co.Bra's 'Planar' indoor TV antenna, which is usable on all UHF TV channels is available from **Banbridge Ltd • 1 York Road • LONDON SW19 8TP.**

Personal Organizer Software

An integrated software package for generating Filofax-compatible output on DAATAfax computer stationery for six-ring bound personal organizers is available from Kempston Data.

The package consists of four modules to produce a diary in a variety of formats; a phonebook data-based with search and sort; and a notepad in free format for storing personal data, timetables, expense sheets, and so on.

Kempston Data Ltd • 21 Linford Forum • Linford Wood • MILTON KEYNES MK14 6LY.

Long-life Memory Back-up

Battery memory back-up packs for business computers that offer real-time clock protection for up to ten years are available from Vertechs Industries. They can be used continuously or when there is a power failure only.

The units are powered by lithium batteries that are smaller and longer-lasting than conventional alkaline-manganese types.

Vertechs Industries Ltd • Unit 3B • Airport Industrial Estate • NEWCASTLE-UPON-TYNE NE3 2EF.

PCB Fault Checker

Faults in electronic assemblies, such as dry solder joints, suspect through-holes, broken tracks and weak crimp joints, can be pinpointed rapidly by a microprocessor-controlled quality checker from Montford Instruments so that failures can

be rectified in the factory and not in the field after delivery.

The 'Quality Checker' uses rapid temperature cycling of 10 °C/min to cause thermal stressing to the PCB under test. During the heating process the components can be power simulated and also monitored by ATE.

Montford Instruments • 24–26 Gorst Road • LONDON NW10 6LE.

Fax and Telex on the Desk

A new concept in business communications that offers not only fax but also telex from the desk, and has the intelligence to decide which of the two message routeings will be the most cost-effective, is available from Trend Telecommunications

Faxtel provides every PC-based word processor in the modern office with direct access to fax and telex machines, regardless of whether the PC is networked or not.

Trend Telecommunications Ltd • Knaves Beech Estate • Loudwater • HIGH WYCOMBE HP10 9QZ

5 V – 5 V LAN Converter

A local area network (LAN) converter that is said to offer high output isolation capability in a compact form to original equipment manufacturers (OEM) of telecommunications systems, digital switching and communications systems and electronic networks has been developed by Gardners Transformers.

The converter provides high-voltage isolation with an input voltage range of 4.75–5.25 V and a nominal output range of 4.9–5.2 V. Output current is 30 mA minimum to 150 mA maximum and switched-mode conversion permits typical line regulations of ±0.5% at an output current of 150 mA, and load regulation of 3% for load changes from 30–100 per cent of full load.

Gardners Transformers Ltd • Somerford Road• CHRISTCHURCH BH23 3PN.

High-integrity Microprocessor

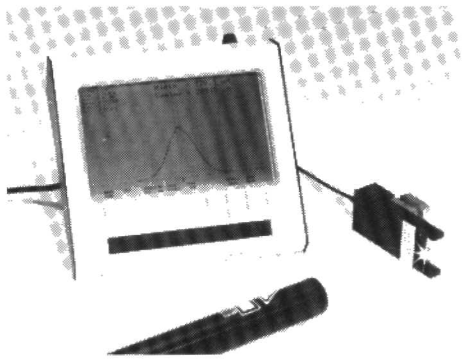
Marconi have designed a 32-bit microprocessor specifically for high-integrity control systems. Called VIPER–1 (verifiable integrated processor for enhanced reliability), it is a gate-level design for use in safety-critical systems, where loss of life, damage to the environment and loss of security are prime considerations.

Marconi Electronic Devices Ltd • IC Division • Doddington Road • LINCOLN LN6 3LF.

NEW PRODUCTS

Optical beam profiler

Instant and safe analysis of laser beams and other optical sources used in modern telecomms, BC and electronics work is provided by DataRay's BeamScope.



The unit employs a single scanning head to cover from UV to IR and will accurately draw the profile of beams with diameters from 1 µm to 25 mm – without adjustment of the head, aperture or calibration.

Lambda Photometrics Ltd • Lambda House • Batford Mill • HARPENDEN AL5 5BZ.

DMAC/D2MAC Tester

The SI7765 DMAC and D2MAC Packet Test Signal Generator from Schlumberger provides comprehensive test pattern and signals to European Broadcasting Union specifications to simplify installation and maintenance of direct broadcast by satellite (DBS) systems. The generator simply replaces ther program for testing a transmission channel, receiver or code converter. The signals are generated by digital synthesis.

Schlumberger Technologies • Instrument Division • Victoria Road • FARNBOROUGH GU14 7PW.

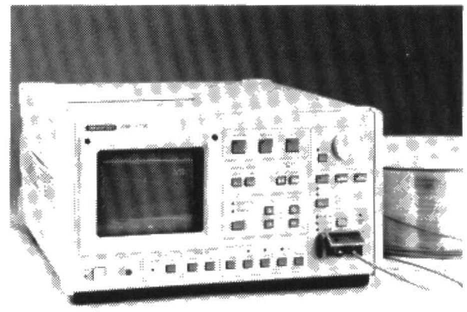
Modular Pushbuttons

The EAO Series 61 family of modular pushbutton switches includes illuminated pushbuttons, keylock switches, illuminated rotary selector switches and indicator lamps. The actuators may be installed totally separately from the switching elements. The devices are available in a variety of sizes.

Highland Electronics (Distribution) Ltd • Albert Drive • BURGESS HILL RH15 9TN.

High-resolution OTDR

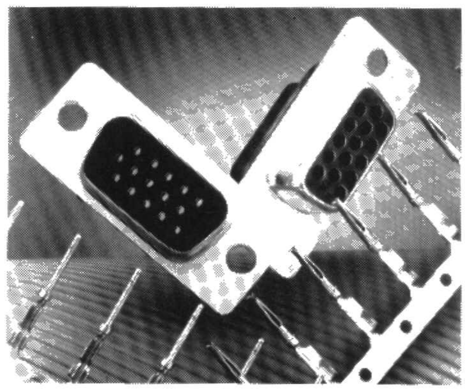
The Advantest range of optical time-domain reflectometers (OTDRs) has been expanded by the Model Q8640. This instrument uses a plug-in concept with two options available at present: 1300 nm single-mode and 1300 nm multi-mode.



Advantest UK Ltd • 10th Floor • CI Tower • St Georges Square • High Street • NEW MALDEN KT3 4HH.

Crimp Connectors from Bulgin

A new range of Beau crimp-insertable D-subminiature connectors, offering 15 contacts in a nine-contact shell is available from Bulgin. These connectors provide 66 per cent more contacts in the same space, thus enabling substantial savings to be made by reducing the cost of installation and replacement.



A.F. Bulgin + Co. • Bypass Road • BARKING IG11 0AZ.

Cordless Soldering Iron

Among the comprehensive range of soldering and desoldering equipment available ex-stock from Axiom Electronics is Weller's Pyropen, a cordless soldering iron that is ideal for field service applications.

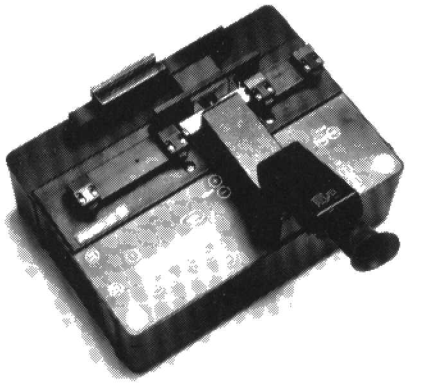
Pyropen is an LP gas-operated iron that weighs only 90 g. The LP gas is stored in the iron's handle and gives up to four hours of continuous operation.

Axiom Electronics • Turnpike Road • Cressex Estate • HIGH WYCOMBE HP12 3NR.

NEW PRODUCTS

Fibre optic cable splicer

The quality of each splice in fibre optic cabling determines to a great extent the transmission quality of the whole network.

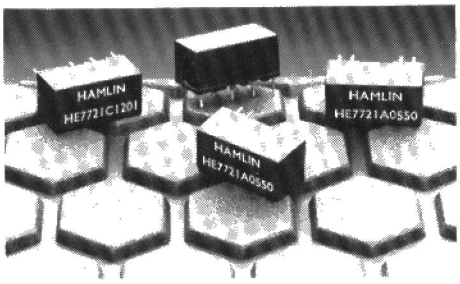


To meet market demands for faster and better splicing – taking less than one minute per joint –Ericsson Cables has added a fourth-generation splicer, the FSU900, to its range of fibre cable splicing equipment.

Ericsson Cables (Att. Inge Berg) • S–172 87 Sundbyberg • SWEDEN.

Sensitive Reed Relays

Hamlin's 7700 Series of sensitive reed relays is available with a choice of dual-in-line (DIL) or electro-mechanical (Type 47) footprints. Their sensitivity is about 40% higher than that of a standard DIL relay: an important factor where power saving is of importance.



Hamlin Electronics • DISS IP22 3AY

Low-noise Preamplifier

The SSM2015 from Precision Monolithics is a differential input instrumentation/microphone amplifier that offers ultra-low-noise performance (1.3 nV/Hz). It is the only instrumentation amplifier available with programmable input stage operating current, which allows noise to be optimized for source impedances of up to 40 kΩ.

Bourns Electronics Ltd • 90 Park St • CAMBERLEY GU15 3NY.

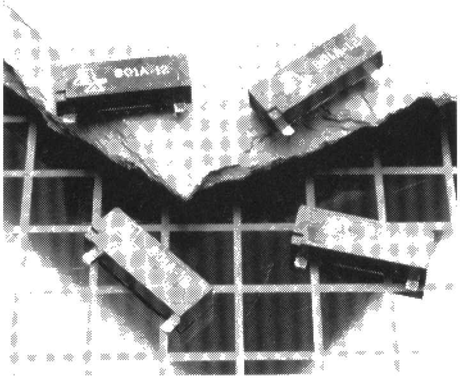
100 MHz Pulse Generator

The microprocessor-based Model 8600 programmable pulse generator can be used manually or via GPIB interface to provide two 5 ns pulses through two independently programmed channels. The pulse period may be set between 10 ns and 2 s, with an amplitude of between 0.5 V and 10 V (pp) within a ±10 V window, and pulse widths can be set between 5 ns and 4s.

Global Specialties • 2nd Floor • 2-10 St. Johns Street • BEDFORD MK42 0DH.

Low-profile DIL Reed Relay

The 800 Series low-profile DIL relay from Astralux Dynamics is designed specifically for use with surface mounting assembly techniques. It is about one third the weight of conventional DIL reed relays and is moulded in high-grade epoxy resin to withstand the temperatures inherent in the vapour phase and wave soldering methods.



Astralux Dynamic Ltd • Red Barn Road • BRIGHTLINGSEA CO7 0SW.

Specialized Electronics Equipment for Secondary Schools

Flight Electronics, already supplying a number of secondary schools throughout the country with electronics equipments, have available a number of instruments that are of particular interest to such schools.

In addition to the popular GOS-935 oscilloscope, there is the GAG-808B audio signal generator, which is proving popular with GCSE students and equipment specifiers alike.

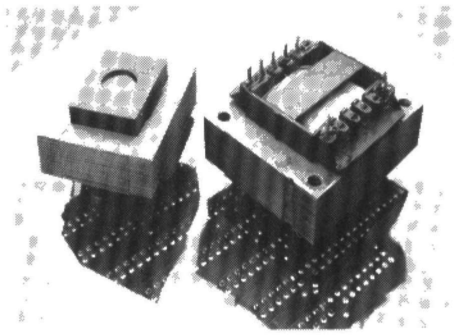
Ideal for introductory electronics work, the GOS-935 is a single-beam instrument with a bandwidth of DC to 5 MHz (vertical) and 500 kHz (horizontal).

The GAG-808B generates signals from 10 Hz to 1 MHz across five ranges. Low distortion, good accuracy and fast rise times are some of its features.

Flight Electronics • Ascupart Street • SOUTHAMPTON SO1 1LU.

PCB Mounting Transformers

Two new PCB mounting transformers with power ratings of up to 6 VA or 15 VA from Iskra have been designed for use in the commercial or professional market for mains, isolating and signalling applications.



The transformers have standard 5 mm pitch pinouts for direct mounting and connection on the board as well as four 3.6 mm dia. holes through the laminations for applications requiring fixing screws.

Special versions for operating at up to 180 °C are available.

Iskra Ltd • Redlands • COULSDON CR3 2HT.

READERS SERVICES

All orders must be sent BY POST to our London office using the appropriate form opposite. Please note that we can not deal with PERSONAL CALLERS, as no stock is carried at the editorial offices. The postal address is given at the back of the form.
All prices shown are net and customers in the UK should add VAT where shown. ALL customers must add postage and packing charges for orders up to £15.00 as follows: UK, £1.00; Europe, £1.50; other countries, £2.00 (surface mail) or £3.00 (airmail). For orders over £15.00, but not exceeding £50.00, these p&p charges should be doubled. For orders over £50.00 in value, p&p charges will be advised.

Software is also available from **TECHNOMATIC LIMITED** (for address, see inside front cover).

In Sweden, printed-circuit boards should be ordered from
ELECTRONIC PRESS
BOX 5505
S-14105 Huddinge
Telephone: 08-710 08 90

SUBSCRIPTIONS

Subscriptions can be provided anywhere in the world: apply to **Elektor Electronics • Worldwide Subscription Service Ltd • Rose Hill • Ticehurst • East Sussex TN5 7AJ • England.**

LETTERS

Letters of a general nature, or expressing an opinion, or concerning a matter of common interest in the field of electronics, should be addressed to The Editor. Their publication in *Elektor Electronics* is at the discretion of the Editor.

PAST ARTICLES

A limited number of past issues can be supplied at the current cover price plus postage & packing as detailed above. If past issues are no longer available, photo copies of the relevant article can always be provided at a price of £1.00 per article plus postage and packing as detailed above.

TECHNICAL QUERIES

Although we are always prepared to assist readers in solving difficulties they may experience with projects that have appeared in *Elektor Electronics* during the PAST THREE YEARS ONLY, we regret that these can not in any circumstances be dealt with by telephone or fax.

COMPONENTS

Components for projects appearing in *Elektor Electronics* are usually available from appropriate advertisers in this magazine. If difficulties in the supply of components are envisaged, a source will normally be advised in the article.

BOOKS

The following books are currently available: these may be ordered from certain electronics retailers or bookshops, or direct from our London office.

301 Circuits	£6.25
302 Circuits	£6.25
303 Circuits	£7.95
Data Sheet Book 2	£8.25
Microprocessor Data book	£8.95

BINDERS

Elektor Electronics binder	£2.95

FRONT PANELS

	No.	Price (£)	VAT (£)
Intelligent time standard	86124-F	15.70	2.36
Autoranging DMM	87099-F	2.80	0.42
Frequency meter	87286-F	10.75	1.61
Microcontroller-driven power supply	880016-F	28.75	4.31
Preamplifier for purists	880132-F	8.25	1.24
Autonomous I/O controller	880184-F	8.50	1.28
Analogue Multimeter	880035-F	7.00	1.05

ORDER FORM

VAT No. 454 135 463

This order should be sent to:

ELEKTOR ELECTRONICS
(PUBLISHING)
DOWN HOUSE
BROOMHILL ROAD
LONDON SW18 4JQ
ENGLAND

Name	
Address	
	Post code
Telephone	Date

Please supply the following. For PCBs, front panels, EPROMs, and cassettes, state the part no. and description; for books, state the full title; for back numbers, state month and year of publication; for photocopies of past articles, state full name of article and month and year of publication. Please use block capitals. For TERMS OF BUSINESS see overleaf.

Qty	No	Description	Price (£)	VAT (£)	Total (£)

Method of payment (tick as appropriate):	Sub-total (£)	
	P&P (£)	
	Total paid (£)	

☐ Bank draft
☐ Cheque (payable to ELEKTOR ELECTRONICS)
☐ Giro transfer (our A/c no. 34 152 3801)
☐ Postal/money order
☐ Access A/c

Card expiry date:

Signature:

SOFTWARE

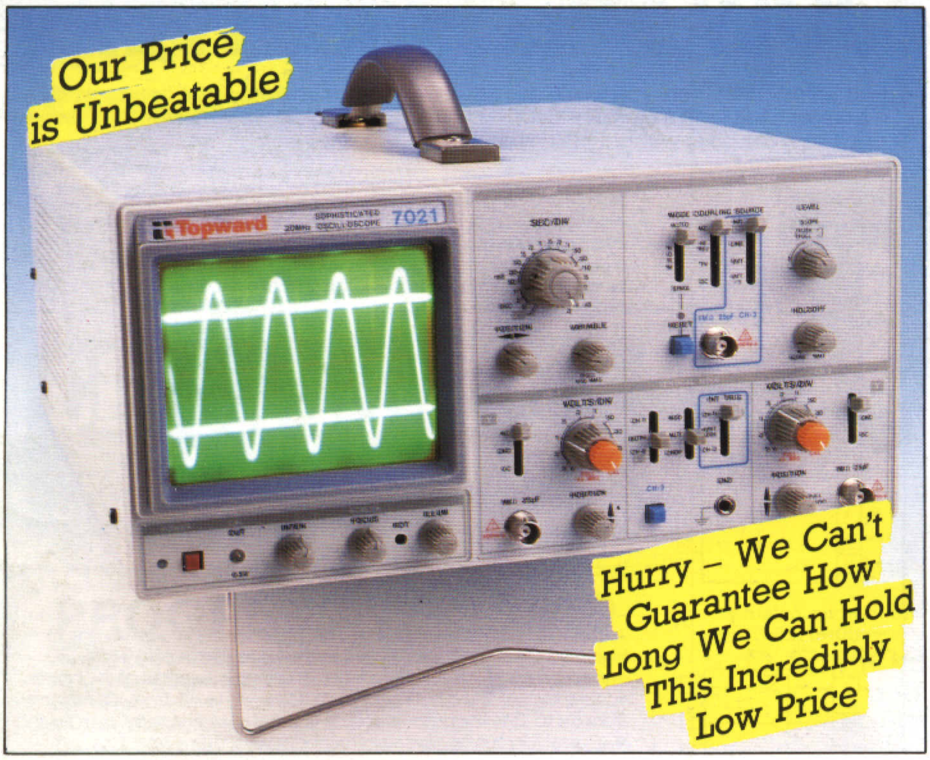
Software in (E)PROMs	No.	Price (£)	VAT (£)
X-Y plotter 1 × 2732	532	9.00	1.35
programmable timer 1 × 2732	535	9.00	1.35
GHz pre-scaler 1 × 2732	536N	9.00	1.35
automate your model railway 1 × 2716	537	7.30	1.10
marine computer 1 × 2716	538	7.30	1.10
Jumbo clock 2 × 2716	539	14.60	2.20
Graphics card 2 × 82S123	543	9.60	1.44
printer buffer 1 × 2716	545	7.30	1.10
MSX EPROMmer 1 × 27128	552UK	7.30	1.10
Intelligent time standard 1 × 2764	553	10.00	1.50
Interface for high-resolution LC screens 1 × 2764	560	10.00	1.50
I/O extension for IBM 1 × PAL16L8	561	8.75	1.32
Centronics interface for slide fader 1 × PAL16R4	562	8.75	1.32
µP-controlled radio synthesizer 1 × 27C64	564	10.00	1.50
Portable MIDI keyboard 1 × 2764	567	10.00	1.50
Pitch control for CD players 1 × 2764	568	10.00	1.50
MIDI control unit 1 × 27C64	570	10.00	1.50
Slave indication unit for I.T.S. 1 × 8748H	700	15.00	2.25
EPROM emulator 1 × 8748H	701	15.00	2.25
Microcontroller-driven power supply 1×8751	702	47.50	7.13
Autonomous I/O controller 1×8751	704	47.50	7.13
Transmission & reception of RTTY	XSS-100	8.00	1.20
(Send formatted 3.5 in diskette containing MSXDOS.COM and COMMAND.COM)			
FAX interface for Atari ST	ESS-102	8.00	1.20
for Archimedes	ESS-103	8.00	1.20
(Send formatted 3.5 in. diskette)			
MIDI keyboard interface	575	10.00	1.50

PRINTED CIRCUITS

Readers who wish to make their own PCBs (for private and personal use only) may in many, but not all, cases receive the relevant drawings free of charge by ordering these on the order form above and enclosing a stamped addressed envelope (preferably 9×6 in or 230×150 mm). These drawings may also be supplied on film (only for projects published during the past three years) priced at £3.00 each plus p&p.

	No.	Price (£)	VAT (£)
NOVEMBER 1988			
Tracker-ball for Atari ST	87260	Not available	
Simplified time-signal receiver	87513	7.40	1.11
Bus interface	880074	16.75	2.51
LFA-150 – a fast power amplifier	880092-1	8.45	1.27
	880092-2	7.70	1.16
Harmonic enhancer	880167	6.30	0.95
Portable MIDI keyboard	880168	7.85	1.18
IR control for stepper motors		Not available	
DECEMBER 1988			
LFA-150: a fast power amplifier	880092-3	6.40	0.96
	890092-4	6.45	0.97
Composite-to-TTL adaptor	880098	4.85	0.73
Colour test pattern generator	880130	13.30	2.00
Autonomous I/O controller	880162	4.50	0.68
	880163	4.85	0.73
	880184	15.30	2.30
Pitch control for CD players	880165	11.50	1.73
JANUARY 1989			
Thyristor speed control	87200	Not available	
Fax interface for Atari ST/Archimedes	880109	7.35	1.10
MIDI control unit	880178-1	9.05	1.36
	880178-2	6.65	1.00
Low-budget capacitance meter	UPB-S1	1.95	0.29
FEBRUARY 1989			
MOSFET power amplifier	87096	12.35	1.85
The digital train	87291-1	4.20	0.63
Touch key organ	886077	10.05	1.16
Car service module	86765	3.60	0.54
	886126	4.10	0.62
VHF receiver	886127	7.45	1.12
Dark-room timer	886100	Not available	
MARCH 1989			
Diesel sound generator	880006	Not available	
VHF/UHF wide-band amplifiers	880186	Not available	
Power line modem	880189	6.10	0.92
ATN Filmnet decoder	890002	Not available	
Centronics buffer	890007-1	19.60	2.94
	890007-2	2.15	0.33
	890007-3	8.35	1.26
APRIL 1989			
The digital train	87291-2/3	4.30	0.65
Class D amplifier	880009	Not available	
DLF frequency reference	880197	Not available	
Function generator	UPBS-1	1.95	0.29
Triplet	890013-1	6.65	1.00
	890013-2	6.80	1.02
Multi-point IR control	890019-1	3.45	0.52
	890019-2	4.05	0.61
Video recording amplifier	889502	See advert on pp.8-9	
MAY 1989			
RDS decoder	880209	Not available	
Digital model train (4)	87291-4	5.25	0.79
Analogue Multimeter	890035	12.50	1.88
In-line RS-232 monitor	890036	Not available	
Code converter for Centronics-compatible printers	890058	Not available	
DTMF system decoder	890060	6.50	0.98
Transistor tester	896029	Not available	
Sine wave converter	UPB-S1	1.95	0.29
Super-VHS to RGB converter		See advert on pp. 8-9	
JUNE 1989			
8-digit frequency meter	880128	11.50	1.73
In-line RS232 monitor	890036	Not available	
In-circuit transistor tester	890079	Not available	
Echo unit		See advert on pp.8-9	
July/August 1989			
MIDI keyboard interface decoder board	890105-1	7.00	1.05
controller board	890105-2	5.25	0.79
Tracking tester		See advert on pp. 8-9	
Test pattern generator	890020	Not available	
DC-AC power converter	890056	Not available	
Floppy disk monitor	890078	4.25	0.64
display board	85019	3.25	0.48
Function generator	UPBS-1	1.95	0.29
Call tone generator	894016	Not available	
Sound level meter	894024	3.50	0.53
Mains-failure indicator	894030	Not available	
Radio beacon converter	894041	Not available	
Low-noise microphone preamplifier	894063	3.75	0.56
Bucket brigade delay line	894055	3.75	0.56

Superb Triple-Trace 20MHz Oscilloscope



Precision laboratory oscilloscope.
3 Channels – 3 Trace.
Sensitive vertical amplifier 1mV/div allows very low level signals to be easily observed.
150mm rectangular CRT has internal graticule to eliminate parallax error.
X-Y mode allows Lissàjous patterns to be produced and phase shift measured.
TV sync separator allows measurement of video signals.
20ns/div sweep rate makes fast signals observable.
Algebraic operation allows sum or difference of Channel 1 and 2 to be displayed.
Stable triggering of both channels even with different frequencies is easy to achieve.
50mV/div output from Ch 1 available to drive external instrument e.g. frequency counter.
A hold-off function permits triggering of complex signals and aperiodic pulse waveforms.

ONLY

£287.49

XJ61R

40MHz Triple-Trace Oscilloscope

As above, but with 40MHz bandwidth and super bright 12kV tube even at the highest frequencies. This instrument also has a delayed sweep time base to provide magnified waveforms and accurate time interval measurement. Truly superb precision instrument.

Includes VAT & Carriage As Well!

ONLY

£499.95

XJ60Q

Order Coupon		Send to P.O. Box 3, Rayleigh, SS6 8LR	
Qty.	Description	Code	Price
		Add carriage	50p
		Total	

Name .... Total

Address ....

.... Post Code ....

I authorise you to debit my Credit Card account for the cost of goods despatched.

Card No. ☐☐☐☐☐☐☐☐☐☐☐☐☐☐☐☐

Access, Amex, Visa .... delete as required.

If ordering by Credit Card please sign ....

Expiry date of Credit Card ....

E

Maplin ELECTRONICS

P.O. Box 3, RAYLEIGH, ESSEX SS6 8LR.

PHONE BEFORE 5PM FOR SAME DAY DESPATCH **0702 554161**

ALL PRICES INCLUDE VAT.

All items subject to availability. Subject to availability both items will be on sale in our shops in Birmingham, Bristol, Leeds, London, Manchester, Nottingham, Southampton and Southend-on-Sea.